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Radio Progress During 1940

THE YEAR 1940 in radio was marked by unusual commercial activity and increasing production as well as engineering advances and the advent of new services for public use.

Year-end statistics compiled by the Institute showed that 11,000,000 radio receivers were produced during 1940, a record-breaking figure. Although in 1940, table-model receivers continued to be the most popular, the increased sale of radio-phonograph console combinations, including in some cases home-recording attachments, was a striking trend. Automobile receivers accounted for 2,300,000 of the receivers made. In all there were approximately 1000 different models of radio receivers available. The number of tubes produced during the year was 110,000,000 another figure never before equaled.

In addition to this large production for public consumption, the industry prepared to produce vast quantities of radio apparatus and parts for the United States Army and Navy defense program. By the end of the year, contracts for such equipment totaling nearly $50,000,000 had been awarded to radio manufacturers.

Trends in Radio Receivers

One of the outstanding developments was the miniature receiver, weighing less than five pounds and powered by small readily renewable batteries. Some 200,000 of them were sold in the latter half of the year. In the larger receivers, the wide adoption of loop antennas, contained within the cabinet, made unnecessary the erection of antennas for local reception.

Automobile radio sets were produced with short-wave tuning bands, making possible direct reception from European stations while traveling. Push-button tuning of automobile sets became widely used.

International Broadcasting

The Federal Communications Commission, in recognition of the increased importance of international broadcasting brought about by the European War, took two steps to increase the effectiveness of American short-wave stations, particularly those serving South America. Regulations were adopted requiring all short-wave broadcast stations in the United States to maintain a minimum power of 50 kilowatts and to employ directive antennas which would further increase the power of the station, in the direction of the intended audience, by a factor of 10. The Commission authorized commercial sponsorship of programs carried by these short-wave stations, thus permitting the broadcaster to regain at least a part of the expense of operating the station. Programs particularly designed for overseas listeners were developed.

Broadcast Stations to Shift Positions

In 1940, the North American Regional Broadcast Agreement signed at Havana in December, 1937, was ratified. Following the provisions of this agreement, the Federal Communications Commission established March 29, 1941, as the date on which the majority of the United States broadcast stations would shift their dial positions, chiefly by small amounts, to permit a more equitable distribution of radio space among the various countries and to reduce interference. Over 90 per cent of the more than 800 standard broadcast stations will be affected.

The year was marked by increases in power by many standard broadcast stations. Many additional broadcast stations, mostly of low power, were also authorized.

Frequency-Modulation Broadcasting Established as Public Service

So far as new services for the public are concerned, an outstanding announcement was made by the Federal Communications Commission in May, 1940, setting aside 40 channels for frequency-modulation high-frequency broadcast stations, five for educational broadcasting purposes and 35 for broadcasting to the public. By the end of 1940, 25 applications for permission to erect frequency-modulation commercial broadcast stations had been granted. Frequency-modulation broadcast stations operating under the new commercial rules (as contrasted with stations operating under experimental authorization) were authorized to charge sponsors for time on the air beginning January 1, 1941, on the same basis as standard broadcast stations.

Frequency modulation was adopted as the medium for two-way police communication in several new installations, including the police systems of the State of Connecticut and the City of Chicago. The use of frequency modulation for aircraft communication was also under investigation.

Television

Plans for commercialization of television broadcasting, which had been scheduled for the end of 1940, were affected by the decision of the Federal Communications Commission to re-examine the question of transmission standards. To carry out this program a large group of experts, the National Television Systems Committee including nine panels or subcommittees, was organized in August and was ready to render its report to the Commission at the end of the year.

Experimental television broadcasting, however, continued in New York, Philadelphia, Schenectady, Chicago, and Los Angeles. In August, 1940, a New York television transmitter concluded fifteen months of

regularly scheduled programs which covered practically every field of entertainment and instruction.

Among the television programs broadcast were notable political events, including the Republican National Convention in Philadelphia, covered by cameras which relayed the program to New York over coaxial cable circuits, and the Democratic National Convention at Chicago covered by motion pictures flown to New York and then broadcast. The major rallies of both parties, held at the end of the campaign in Madison Square Garden, New York, brought both presidential candidates before the television cameras.

Among the important technical developments were the use of ordinary telephone circuits for relaying television programs over short distances, the development of a radio-relay system operating on wavelengths of less than 1 meter, and the demonstration of television reception in full color. Larger pictures were shown both by the use of larger picture tubes (up to 20 inches in diameter) and by projection on a screen measuring $4\frac{1}{2}$ by 6 feet. An inexpensive television camera tube was announced for amateur experimenters.

Facsimile

Facsimile picture transmission as a broadcast service to the public was not radically extended, but its use in newspaper work increased considerably, particularly in transmitting war pictures from Europe. Facsimile service in wire message telegraph service was extended to include the Atlanta and San Francisco areas.

General

This summary of progress during 1940 covers, in general, the period up to the first of November. It is based on material prepared by members of the 1940 Annual Review Committee of the Institute of Radio Engineers with the collaboration of the Institute's 1940 Public Relations Committee. The final editing and coordinating of the material and the preparation of the introductory section in behalf of the Annual Review Committee was carried out by Laurens E. Whittemore, chairman; Harold A. Wheeler, and Keith Henney, with John D. Crawford acting as secretary.

The individual reports on the special fields were prepared by the following chairmen of the Institute's 1940 technical committees.

P. T. Weeks, Technical Committee on Electronics
E. G. Ports, Technical Committee on Transmitters and Antennas
D. E. Foster, Technical Committee on Radio Receivers
D. E. Noble, Technical Committee on Frequency Modulation
I. J. Kaar, Technical Committee on Television
J. L. Callahan, Technical Committee on Facsimile
H. S. Knowles, Technical Committee on Electro-acoustics
J. H. Dellinger, Technical Committee on Wave Propagation

The chairmen of the above committees wish to acknowledge the assistance given by their individual members.

PART I—ELECTRONICS*

Cathode-Ray and Television Tubes—High-Vacuum Transmitting Tubes—Gas-Filled Tubes—Ultra-High-Frequency Tubes—Small High-Vacuum Tubes—Photoelectric Devices

Cathode-Ray and Television Tubes

Until recently, developments in cathode-ray devices had to do mostly with tubes for oscilloscopic purposes and for the pickup and reproduction of television signals. This year the literature in this field dropped off somewhat, probably because of the slackening of television activities, and we find an increasing number of papers dealing with various aspects of the electron microscope.

Some activity has continued, of course, in all phases of cathode-ray-tube development. One interesting variation in fluorescent-screen preparation is the mounting of the phosphor on a thin metal foil through which the electron beam must pass before striking the phosphor. The customary sticking potential can thus be avoided completely and certain other advantages obtained. Sufficient velocity must be attained by the electrons to penetrate the metal film and still strike the phosphor with the desired energy.

Some of the older problems of the art are still being studied, namely deflection and focusing distortions and the factors affecting improved electron-beam control.

Light valves actuated by electron beams and suitable for television purposes are based on controlled migration of "color centers" in certain types of crystals while another makes use of controlled orientation of particles suspended in a liquid. Both are transmission-type devices that are reported to have had actual application to image reproduction.

An increasingly large number of papers has appeared dealing with electron microscopes and the accompanying problems. Although transmission-type microscopes have been developed using both the electrostatic and electromagnetic systems, the magnetic type apparently has received the greater emphasis. The literature dealing with magnetic-lens distortions is growing and one commercial electron microscope using magnetic lenses has been placed on the market in this country. Stereoscopic photography has been undertaken with some success using an electrostatic electron microscope.

* Decimal classification: R330×621.375.1.

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(2) M. von Ardenne, "The electron microoscillograph," *Hochfrequenz. und Elektroakustik*, vol. 54, pp. 181–188; December, 1939.
(3) H. E. Hollmann, "Microwave oscillography," *Hochfrequenz. und Elektroakustik*, vol. 54, pp. 188–190; December, 1939.
(4) B. C. Fleming-Williams, "Trapezium distortion in cathode ray tubes," *Wireless Eng.*, vol. 17, pp. 61–64; February, 1940.
(5) H. Becker and A. Wallraff, "Astigmatism of magnetic lenses," *Arch. für Elektrotech.*, vol. 34, pp. 43–48; January, 1940.
(6) H. Becker and A. Wallraff, "Image error investigation on a magnetic lens free from image rotation," *Arch. für Elektrotech.*, vol. 34, pp. 115–120; February, 1940.
(7) M. von Ardenne, "Results with a new electron supermicroscope equipment," *Naturwiss.*, vol. 28, pp. 113–127; February, 1940, and vol. 28, pp. 248–252; April, 1940.
(8) H. Mahl, "Stereoscopic photographs with the electrostatic supermicroscope," *Naturwiss.*, vol. 28, p. 264; April, 1940.
(9) H. Becker and A. Wallraff, "Measurement of image errors in an electron optical lens enclosed in iron with a variable air slit," *Arch. für Elektrotech.*, vol. 34, pp. 230–236; April, 1940.
(10) W. Reusse and N. Ripper, "The influence of space-charge on the focusing of cathode-ray tubes," *Teleg.- Fern.- und Funk-Tech.*, vol. 29, pp. 68–74; March, 1940.
(11) J. R. Pierce, "Rectilinear electron flow in beams," *Jour. Appl. Phys.*, vol. 77, pp. 548–554; August, 1940.
(12) H. Kalden, "Investigations with the super microscope," *Chem. Ztg.*, vol. 64, pp. 129–133; 1940.
(13) L. Marton, M. C. Banca, and I. F. Bender, "A new electron microscope," *RCA Rev.*, vol. 5, pp. 232–243; October, 1940.
(14) Eleven articles dealing with all phases of electron microscopy, *Jahr. der AEG Forschung*, Vol. 7, March, 1940.

High-Vacuum Transmitting Tubes

During the year 1940 there was progress in the development of air cooling of large tubes. The conditions of heat flow in the air coolers made of radiating copper fins attached to the anode were investigated theoretically. Such studies materially aided in the development of more efficient coolers. Power tubes for the generation of power at very high frequencies have received considerable attention, but progress in this field is reported in the section on ultra-high-frequency tubes. Attention is called to a number of papers dealing with the properties of electron flow in vacuum tubes both from a theoretical and experimental point of view and contributions to the methods of testing power tubes.

(1) E. M. Ostlund, "Air cooling applied to external-anode tubes," *Electronics*, vol. 13, pp. 36–39; June, 1940.
(2) I. E. Mouromtseff, "The influence of 'cold ends' in the performance of vacuum-tube air coolers," *Phys. Rev.*, vol. 58, p. 204; July, 1940. (Summary only.)
(3) R. Warnecke and R. Bonne, "High-power transmitting triode for short waves," *L'Onde Élec.*, vol. 19, pp. 39–47; January-February, 1940.
(4) K. Spangenberg, "Current division in plane-electrode triodes," Proc. I.R.E., vol. 28, pp. 226–236; May, 1940.
(5) G. Stolzer and J. A. Sargrove, "Measurement of transmitting valve characteristics above the dissipation limit," *Electronics and Telev.*, vol. 13, pp. 153–156; April, 1940.
(6) O. W. Livingston, "Oscillographic method of measuring positive-grid characteristics," Proc. I.R.E., vol. 28, pp. 267–268; June, 1940.
(7) I. E. Mouromtseff, "Water and forced-air cooling of vacuum tubes with external anode," Proc. I.R.E., vol. 28, p. 251; May, 1940. (Abstract only.)
(8) I. E. Mouromtseff and W. G. Moran, "Large air-cooled tubes in 50 kilowatt transmitters," Proc. I.R.E., vol. 28, p. 251; May, 1940. (Abstract only.)

Gas-Filled Tubes

A new type of mercury-pool rectifier was described which, it is claimed, has been used for rectifying voltages as high as 105,000 volts.

(1) T. M. Libby, "Surface-controlled mercury-pool rectifier," Proc. I.R.E., vol. 28, pp. 52–59; February, 1940.

Ultra-High-Frequency Tubes

There was continued activity in this field covering a wide variety of devices. Advances were made in both theory and development, but the published papers were devoted chiefly to refinement of technique and more searching analysis of devices and features previously discussed. Such things as the input loading of "space-charge control tubes" and the circuit impedance properties of "cavity resonators" may be mentioned as typical examples of the trend in these publications, and very little of an outstanding nature can be selected as representing a distinct advance during the year 1940.

The situation is covered in outline form by the bibliography which is grouped broadly under headings representing the major lines of activity in the field.

Space-Charge Control Tubes (Negative Grid)

(1) H. Affolter, "Demontierbare Senderöhren für grosse Leistung (High-power transmitting tubes which can be demounted," *Tech. Mitteilungen*, vol. 18, pp. 49–51; April 1, 1940.
(2) "Demountable high-vacuum tubes," Proc. I.R.E., vol. 28, pp. ii, iv; April, 1940.
(3) G. P. Harnwell, and L. N. Ridenour, "Vacuum tubes in the physics laboratory," *Amer. Jour. Phys.*, vol. 8, pp. 79–99; April, 1940.
(4) R. King, "Continuously variable oscillator for parallel line measurements at 100 to 1000 megacycles." *Rev. Sci. Instr.*, vol. 11, pp. 270–271; August, 1940.
(5) R. King, "Parasitic electronic oscillations and coupling frequencies in a power tube," *Jour. Appl. Phys.*, vol. 11, pp. 615–620; September, 1940.
(6) A. Marino, "Sulla teoria dell'amplificazione a radiofrequenza (On the theory of radio-frequency amplification)," *Alta Frequenza*, vol. 9, pp. 67–102; February, 1940.
(7) K. Morita, "Method of loading micro-wave generators," *Nippon Elec. Commun. Eng.*, no. 19, pp. 182–187; January, 1940.
(8) B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuations in space-charge-limited currents at moderately high frequencies. Part II—Diodes and negative-grid triodes," *RCA Rev.*, vol. 4, pp. 441–472; April, 1940; vol. 5, pp. 106–124; July, 1940.
(9) K. Posthumus, and C. A. Gehrels, "Some typical phenomena relating to ultra-short-wave pentode amplifier circuits," *Philips Transmitting News*, vol. 6, pp. 49–55; December, 1939.
(10) F. Preisach and I. Zackarias, "Input conductance measurement in high-slope H-F amplifier valves," *Wireless Eng.*, vol. 17, pp. 147–157; April, 1940.
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(12) G. Rodwin and L. M. Klenk, "High-gain amplifier for 150 megacycles," Proc. I.R.E., vol. 28, pp. 257–261; June, 1940.
(13) H. Rothe, "Operation of electron tubes at high frequencies," Proc. I.R.E., vol. 28, pp. 325–331; July, 1940.
(14) S. W. Seeley and E. I. Anderson, "U-H-F oscillator frequency-stability considerations," *RCA Rev.*, vol. 5, pp. 77–88; July, 1940.
(15) N. Tanaka, "Determining the limit of stable operation for the short-wave high power amplifiers," *Nippon Elec. Commun. Eng.*, no. 19, pp. 148–153; January, 1940.
(16) N. Tanaka, "Inverted type short wave high power amplifier," *Nippon Elec. Commun. Eng.*, no. 19, p. 192; January, 1940.
(17) "Transmitting tube," *Electronics*, vol. 13, p. 75; May, 1940.
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Positive-Grid Tubes

(22) W. S. Elliott and J. A. Ratcliffe, "Barkhausen-Kurz oscillations with positive ions (Letter to the editors)," *Nature*, vol. 145, pp. 265–266; February 17, 1940.

(23) R. Golicke, "Schwingungserzeuger mit strombegrenzten Mehrgitterröhren (Oscillation generators with current-limited multigrid tubes)," *Elek. Nach. Tech.*, vol. 16, pp. 286–297; November–December, 1939.

(24) H. H. Klinger, "Zur Gleichrichtung und Entdämpfung ultrakurzer Wellen mit dem Bremsaudion (On the detection and improvement of ultra-short waves with the retarded field audion)," *Funk-Tech. Monatschefte*, no. 7, pp. 101–106; July, 1940.

Magnetrons

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(31) E. B. Moullin, "Considerations of the effect of space-charge in the magnetron," *Proc. Camb. Phil. Soc.*, vol. 36, pp. 94–100; January, 1940.

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(33) A. Okazaki, "Induced currents in split cylindrical electrodes by moving charges," *Electrotech. Jour.* (Tokyo), vol. 4, pp. 46–47; February, 1940, pp. 83–87 and 87–90; April, 1940.

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(35) K. Owaki, "On the magnetron with a bowl type resonator," *Electrotech. Jour.* (Tokyo), vol. 4, p. 188; August, 1940.

(36) K. Owaki, "Partial split anode magnetron," *Nippon Elec. Commun. Eng.*, no. 19, pp. 188–191; January, 1940.

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(40) H. Döring and L. Mayer, "Geschwindigkeitsgesteuerte Laufzeitröhren (Velocity-controlled transit-time tubes)," *Elek. Nach. Tech.*, vol. 61, pp. 685–690; July 25; pp. 713–715; August 1, 1940.

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Small High-Vacuum Tubes

The trend of late 1939 in reducing the number of new tube types announced continued through the year 1940. Only one third as many types were announced in 1940 as in 1939. Of the new ones none involved new fundamental principles of electric design. In general, the majority of these were lock-in and GT (small glass envelope and octal-base) constructions having performance characteristics similar to those of existing types.

The year witnessed a record number of tubes sold. Efforts by tube manufacturers to concentrate demand on preferred types made substantial progress. This move is resulting in larger production volume on a relatively few tubes with consequent cost and quality advantages for these types. There also was an industry movement to eliminate some G (large glass envelope and octal base) types by combining them with GT types under a double type-number designation.

A new method of manufacturing metal tubes was described in which the degassing of the tube parts and the sealing-on of the envelope are combined in an operation carried out in a vacuum under a bell jar. No seal-off tip is required with this method.

(1) D. W. Jenks, "The evolution of a new type of metal receiving tube," *Electronics*, vol. 13, pp. 26–27; December, 1940. (Abstract only.)

Photoelectric Devices

There was comparatively little activity during 1940 on photoelectric devices beyond a continued expansion in their practical application. A new phototube having sensitivity to blue light and the near-ultraviolet became commercially available. For many uses this new phototube is superior to cells with the caesium-oxygen-silver surface. Another phototube, employing a 9-stage electrically focused electron multiplier from which a multiplication of 5 per stage is obtainable, was described.

A report was published on the important contribution of metastable atoms to the time lag found in gas-filled photoelectric cells.

Continued research in the manufacture of barrier-layer photocells has resulted in material improvements over the past year. Such improved cells are now available and these have a more nearly linear output with materially increased stability.

(1) A. M. Glover and R. B. James, "A new high-sensitivity photosurface," *Electronics*, vol. 13, pp. 26–27; August, 1940.
(2) R. W. Engstrom and W. S. Huxford, "Time-lag analysis of the Townsend discharge in argon with activated caesium electrodes," *Phys. Rev.*, vol. 58, pp. 67–77; July 1, 1940.
(3) J. A. Rajchman and R. L. Snyder, "An electrically-focused multiplier phototube," *Electronics*, vol. 13, pp. 20–23, 58, 60; December, 1940.

PART II—RADIO TRANSMITTERS AND TRANSMITTING ANTENNAS*

Transmitters—Transmitting Antennas—Navigational-Aid Systems—International Broadcast Programs—Broadcast Regulation

Transmitters

In the broadcast transmitter field there is little to report except an increase in the trend mentioned last year toward the use of air cooling for the higher-powered transmitters, making available such broadcast transmitters with outputs as great as 50 kilowatts. There has been a considerable increase in the number of domestic broadcast stations in the United States during 1940† and also an increase in the average power of quite a number of these stations; many which formerly operated on 100 watts have increased to 250 watts and a number have changed from the 5000-watt class to the 50,000-watt class.

Unsettled conditions throughout the world have placed added importance on international broadcasting. This year, for the first time, the Federal Communications Commission has licensed international broadcast stations to operate on a commercial basis. Commercialization has provided a stimulus to this service and it may continue to do so to an increasing degree.

Effective July 1, 1940‡, new rules of the Federal Communications Commission made it compulsory for each licensed international broadcast station to operate with a minimum transmitter power of 50 kilowatts and a minimum directive antenna power gain of 10. As a result, all such stations in the United States have built, or are building, their facilities to meet or exceed these requirements. The trend in the design of international transmitters has been toward the use of high-level modulating systems utilizing class B modulators.

The year 1940 saw the adoption by the Federal Communications Commission of rules looking toward the commercialization of high-frequency broadcasting by stations employing frequency modulation (FM). The first permits authorizing the construction of commercial stations were granted by the Commission during October, 1940, and many additional ones were granted before the end of the year.§ Commercial operation was permitted after January 1, 1941, on the part of stations complying with the new rules, although the importance of the new service as an advertising medium will depend upon the rapidity with which the public purchases receivers capable of receiving frequency-modulated signals. Transmitter designs have been completed by several manufacturers and there are now available frequency-modulation transmitters ranging in power from 50 watts up to 50,000 watts.

In the communication field, transmitters employing air-cooled tubes with outputs up to approximately 5000 watts have been manufactured by several organizations for use in the police, airways, and point-to-point services. These transmitters are characterized by complete duplication of the radio-frequency tubes and their associated circuits, thus avoiding complications incident to radio-frequency switching.

National defense requirements have resulted in the purchase of substantial quantities of equipment by the various government departments. Very considerable numbers of transmitting equipments for all fields of application are being delivered by the manufacturers or are now under contract. The design of one of the transmitters now on order by a nonmilitary government agency is unique in providing for 20 kilowatts output at frequencies up to 24 megacycles with all air-cooled tubes.

A number of papers relating to transmitters have appeared which are of interest.

(1) C. E. Strong, "The inverted amplifier," *Electronics*, vol. 13, pp 14–16, 55–56, July, 1940.
(2) F. H. Kroger, Bertram Trevor, and J. Ernest Smith, "A 500-megacycle radio-relay distribution system for television," *RCA Rev*, vol 5, pp 31–50, July, 1940
(3) Andrew V. Haeff and Leon S. Nergaard, "A wide-band inducttive-output amplifier," Proc. I.R E., vol. 28, pp. 126–130; March, 1940
(4) Arthur W. Melloh "Damped electromagnetic waves in hollow metal pipes," Proc I R E vol 28, pp. 179–130; April, 1940.
(5) W. L. Barrow and W. W Mieher "Natural oscillations of electrical cavity resonators," Proc I R E., vol. 28, pp. 184–191; April, 1940
(6) W. P. Mason, "A new quartz-crystal plate, designated the GT, which produces a very constant frequency over a wide temperature range" Proc I.R E, vol 28, pp. 220–223; May, 1940.

Transmitting Antennas

The use of directional antennas increased rapidly for American broadcast stations. By September of 1940, there were 116 directional antennas in use by United States stations in the standard broadcast band, and this figure probably approximates 140 as of the end of the year. At least 50 additional directional antennas will be required under the terms of the Havana Treaty of December, 1937. The majority of these directional antennas are employed at regional stations and are designed so that the station's nighttime power can be increased from 1000 to 5000 watts without increasing the interference to other stations operating on the same frequency. However, at least 11 50,000-watt transmitters are now using directional antennas. It is inter-

* Decimal classification: R350×R320.

† On January 1, 1941, the number of licensed standard broadcast stations in the United States was 831; construction permits had been issued by the Federal Communication Commission for 51 additional stations.

‡ Effective date of requirement as to transmitter power later postponed until July 1, 1941.

§ See page 97, footnote †.

esting to note that many of these antennas involve increasingly elaborate systems utilizing three or four radiating elements and that vertical as well as horizontal directivity is considered.

Further work has been done on horn radiators, particularly with reference to multiunit systems.

(1) Andrew Alford and A. G. Kandoian, "Ultra-high frequency loop antennas," *Trans. A.I.E.E.*, vol. 59, pp. 843–848; 1940.
(2) W. E. Jackson, A. Alford, P. F. Byrne, H. B. Fischer, "The development of the Civil Aeronautics Authority instrument landing system at Indianapolis," *Trans. A.I.E.E.*, vol. 59, pp. 849–858; 1940.
(3) W. L. Barrow and Carl Shulman, "Multiunit electromagnetic horns," Proc. I.R.E., vol. 28, pp. 130–136; March, 1940.

Navigational-Aid Systems

A horizontally polarized loop transmitting antenna was introduced in the application of directivity to the design of instrument landing systems to avoid the effects of local reflections.

During the year, the Civil Aeronautics Administration of the United States Government has programmed the installation of a number of instrument landing equipments at representative airports throughout the United States in furtherance of their extension of additional navigational-aid systems to aircraft. The same body also planned the installation of a considerable number of 20-kilowatt high-frequency transmitters for transoceanic aids. Considerable work was done in the utilization of ultra-high frequencies in aerial navigation.

Of interest was a trial application of radio (in operation for a few months) as an aid for the motorist at congested places such as on the George Washington Bridge in New York wherein by tuning to 550 kilocycles the motorist was given instructions regarding exits, special traffic conditions, etc. The field was confined to the territory immediately surrounding the bridge itself.

Radio contributed further to the collection of weather data through the use of automatic weather stations at isolated locations, such as mountain peaks, small islands, etc.

(1) W. E. Jackson, "The impetus which aviation has given to the application of ultra-high frequencies," Proc. I.R.E., vol. 28, pp. 49–51; February, 1940.
(2) "Tune 550—Highway radio ahead," *Electronics*, vol. 13, p. 32; September, 1940.
(3) Harry Diamond and Wilbur S. Hinman, Jr., "An automatic weather station," *Nat. Bur. Stand. Jour. Res.*, vol. 25, pp. 133–148; August, 1940 (RP1318).
(4) K. O. Lange. C. B. Pear, and T. A. Dickey, "An automatic radio weather station," *Bull. Amer. Meteorol. Soc.*, vol. 21, pp. 76–77, March, 1940.

International Broadcast Programs

The intense interest in the international situation resulted in thousands of broadcasts over the American networks of pickups from foreign countries. The radio listener can thus expect several times a day to hear the latest news as broadcast from the capitals of the various warring countries. Transmission of programs from the United States for foreign consumption averages 16 hours per day for the various high-frequency broadcasters. One of the broadcast companies states that programs are presented in six different languages.

Broadcast Regulation

A date (March 29, 1941) was set for the reallocation of approximately 90 per cent of more than 800 United States broadcast stations within the frequency band 550–1600 kilocycles. This shift of the United States stations is a part of the procedure involved in putting into practice the provisions of the North American Regional Broadcast Agreement signed in Havana in December, 1937.

(1) The following mimeographed material was issued on September 11, 1940 by the Federal Communications Commission with respect to the proposed reallocation of broadcast stations as of March 29, 1941, pursuant to the North American Regional Broadcast Agreement.
 (a) Public Notice, No. 43249.
 (b) North American regional broadcast agreement—proposed reallocations and their effect, No. 43243.
 (c) Order adopting amendments to rules and regulations of F.C.C. governing standard broadcast station, No. 43248.
 (d) List, by frequencies, of proposed assignments of United States standard broadcast stations, No. 43242.
 (e) List, by call letters, of proposed assignments of United States standard broadcast stations, No. 43252.
(2) "Directory of broadcasting stations of the United States and Canada," *Broadcasting*, vol. 20, "1941 Yearbook Number," pp. 103–183; February 1, 1941. A list, by states and provinces, of frequency assignments for standard broadcast stations pursuant to the North American Regional Broadcast Agreement.
(3) "Log of Canadian broadcast stations, effective March 29, 1941," *Broadcasting*, vol. 19, p. 17; October 15, 1940. A list by call letters.

PART III—RADIO RECEIVERS*

Broadcast Receivers—Circuit Trends—Communication and Navigational-Aid Receivers—Statistics on 1940 Broadcast Receivers

Broadcast Receivers

There appeared on the market several models of really portable receivers much more compact than heretofore but still having excellent sensitivity. To secure minimum size, miniature-type tubes were used in many cases and the size of components was radically reduced.

Many models of broadcast receivers incorporated a range for the reception of frequency-modulated waves.

Short-wave "spread bands," with the principal broadcast high-frequency ranges each occupying the full dial scale, became widely used after having been virtually abandoned for two or three years.

Short-wave ranges were added to automobile receiving sets for the first time, these ranges being almost universally of the spread-band type.

* Decimal classification: R360.

Another general trend was toward an increase in phonograph-radio-combination models. This trend was aided by a general reduction in the price of records during the year. Many more types of automatic record-changers were available than formerly and were generally of improved design.

Home-recording means was a part of many more phonograph models than when it was first revived in 1939.

Phonograph-combination receivers appeared on the market using a photoelectric cell and light source excited from the receiver oscillator. The light beam was reflected from the source to the cell by a small mirror on the end of the needle.

Circuit Trends

Increased attention was paid to improvements in loop-antenna design in order to secure better sensitivity and to intermediate-transformer design in order to obtain high gain. Fluctuation noise in receivers also had more attention than heretofore.

Oscillator-frequency stability, particularly for ultra-high-frequency ranges, was greatly increased by the use of techniques formerly known but not generally appreciated or applied.

While the use of inverse feedback continued, it was not on as wide a scale as formerly. There was more attention paid to the fundamental relationship between phase shift and attenuation in such circuits.

Communication and Navigational-Aid Receivers

In communication receivers improved types of noise limiters were widely used for reducing electrical interference and greatly improved intermediate-frequency-amplifier selectivity characteristics were secured through the use of wave-filter sections in place of simple tuned circuits.

Radio is so important to aviation that there has been a steady increase in the uses to which it has been put in that service and a continuing development of receiving equipment for this purpose. It is used for communication, radio range, and marker-beam applications.

Tests were also conducted with a 63-megacycle radio range system. All of these added uses increased the equipment required on the airplane so that apparatus redesign occurred to incorporate the added services without increase in size or weight.

It came to be realized that one of the fundamental limitations to extending the range of the ultra-high-frequency transmission is due to the inherent tube and circuit noise in receiving equipment. Considerable progress was made in methods of designing receiving sets to minimize this disturbing noise, both by designing new tubes and by using new circuits.

Statistics on 1940 Broadcast Receivers

Models Offered for Sale

Total number of models—approximately 1000
Average number of tubes (all models) 6.8 per model

Average List Price of Models Offered for Sale

All models	$ 76.60
Radio-phonograph-recorder models	141.00
Radio-phonograph models	133.00
Console models	89.00
Table models	31.80
Portable models	29.00

Estimated Manufacturing Quantities by Types

Table models	5,700,000
Automobile receivers	2,300,000
Portable models	1,300,000
Console models	900,000
Radio-phonograph-combination models	800,000

Manufactured Quantities

It appeared virtually certain that a new high would be reached in quantities of both receivers and tubes manufactured, although the increase was by no means as great as occurred in 1939. Probable figures for the year 1940 were

Receivers	11,000,000
Tubes	110,000,000

(1) Martin Wald, "Noise suppression by means of amplitude limiters," *Wireless Eng.*, vol. 17, pp. 432–438; October, 1940.
(2) "Valve and circuit noises," *Wireless World*, vol. 46, pp. 262–265; May, 1940.
(3) B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuations in space-charge-limited currents at moderately high frequencies," *RCA Rev.*, part I, vol. 4, pp. 269–285; January, 1940; part II, vol. 4, pp. 441–472; April, 1940; part III, vol. 5, pp. 106–124; July, 1940; part IV, vol. 5, pp. 244–260; October, 1940; part V, vol. 5, pp. 371–388; January, 1941.
(4) S. Ballantine, "Grid induction noise in vacuum tubes at ultra-high frequencies," Proc. I.R.E., vol. 48, pp. 143–144; March, 1940. (Abstract only.)
(5) R. I. Kinross, "Reducing interference," *Wireless World*, vol. 46, pp. 382–385; September; pp. 432–436; October; pp. 469–470; November, 1940.
(6) "Noise limiter," *Wireless World*, vol. 46, p. 427; October, 1940.
(7) R. G. Herzog, "Short-wave auto radio," *Communications*, vol. 20, p. 9; July, 1940.
(8) W. E. Jackson, "The impetus which aviation has given to the application of ultra-high frequencies," Proc. I.R.E., vol. 28, pp. 49–51; February, 1940.
(9) S. W. Seeley and E. I. Anderson, "UHF oscillator frequency stability considerations," *RCA Rev.*, vol. 5, pp. 77–88; July, 1940.
(10) W. van B. Roberts, "The limits of inherent frequency stability," *RCA Rev.*, vol. 4, pp. 478–484; April, 1940.
(11) J. B. Moore and H. A. Moore, "I-F selectivity in receivers for commercial radio services," vol. 4, pp. 319–343; January, 1940.
(12) H. W. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421–454; July, 1940.

PART IV—FREQUENCY MODULATION*

On May 22, 1940, the Federal Communications Commission issued its order No. 67, designating 40 channels each 200 kilocycles wide in the frequency band 42,000 to 50,000 kilocycles for use by high-frequency frequency-modulation (FM) broadcast stations. Under this order, 5 channels are set aside for use by educational broadcast stations and 35 channels are assigned to commercial broadcasting.†

Two types of frequency-modulation transmitters are on the market. One type produces phase modulation by combining the modulation sidebands from a balanced modulator with the carrier whose phase has been shifted 90 degrees with reference to the usual carrier-to-sideband phase relationship found in an amplitude-modulated wave. Multipliers are used to increase the phase deviation produced by the phase modulator. The carrier frequency is directly controlled by a crystal oscillator.

In the second type of transmitter, frequency modulation is produced by a reactance-tube-controlled oscillator. In one version of this type of transmitter the carrier frequency of the oscillator is controlled by a voltage derived from a converter and discriminator circuit where the converter input voltages are supplied by the output of the transmitter and of a crystal-controlled oscillator. In a second transmitter of the reactance-tube-modulator type, the carrier frequency is synchronized with the output of a reference crystal oscillator by means of an automatically controlled motor-driven variable condenser.

The frequency-modulation broadcast receivers developed during 1940 follow the conventional pattern of the wide-band-pass superheterodyne with a limiter and frequency-detector system. Several companies are now manufacturing combination broadcast receivers for receiving from amplitude-modulation stations in the standard broadcast band and from frequency-modulation stations in the high-frequency band.

The Connecticut State Police completed in 1940 a state-wide two-way frequency-modulation radiotelephone system for all communication with mobile units. Frequency-modulation stations also came into use for this class of service in Chicago and some other localities. The Connecticut system makes use of ten 250-watt crystal-controlled frequency-modulation transmitters operating on a frequency of 39.5 megacycles. Some 225 mobile stations operate 25-watt frequency-modulation transmitters on a frequency of 39.18 megacycles. It is reported that complete state-wide coverage is achieved by the system with a good factor of safety.

* Decimal classification: R414.

† Construction permits issued by the Commission to 25 commercial frequency-modulation broadcast stations were outstanding on January 1, 1941. No commercial licenses under the new rules had been issued by the Commission as of this date but operating authorizations were outstanding to more than 20 experimental frequency-modulation stations.

(1) Samuel Sabaroff, "System of phase and frequency modulation," *Communications*, vol. 20, pp. 11–12 and 24; October, 1940.
(2) W. R. David, "F-M broadcast transmitters," *Communications*, vol. 20, pp. 8–9; October, 1940.
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PART V—TELEVISION*

Programs—Technical Developments—Test Equipment—Amateur Operations—Standards and System Analysis—Applications

The year 1940 witnessed substantial and significant advances in television.

Programs

During the year experimental television broadcasting was continued in various parts of the country† including the New York City, Los Angeles, Philadelphia, Chicago, and Schenectady areas. The greatest activity of this kind was in the New York City area, where at the close of the year, it was estimated there were approximately 4000 home receivers in use.

At the beginning of August, 1940, the television transmitter at the Empire State Building in New York completed 15 months of regularly scheduled programs, given a field test of great value for the future development of television, both in its technical aspects and in program technique. Programs included the broadcasting of plays, variety programs, on-the-spot news pick-ups, movies, indoor and outdoor athletic events, and other subjects.

A special feature in this series was the televising of an airplane view of New York from the sky. This was relayed from the plane to the Empire State Building transmitter for rebroadcasting, and was particularly interesting because of its military possibilities. Another program milestone was the broadcasting of the national political conventions.

Technical Developments

Important developments in program relays, larger pictures, and color television featured the technical advances of 1940. At the Helderberg transmitter near Schenectady, N. Y., the programs from the Empire State Building transmitter in New York City were received and rebroadcast to the surrounding area at a different carrier frequency. The quality of these programs was excellent despite the 129-mile distance between the two transmitters. Equally interesting was the radio relay between the Empire State Building and Riverhead, L. I. In this case radio repeating stations were located at two intermediate points, Hauppauge and Rocky Point, L. I. There was no demodulation at the intermediate points, only amplification and change of carrier frequency. The relay carrier frequencies were around 500 megacycles. High-gain directive antennas were used.

During the year carrier transmission of television signals was demonstrated over the coaxial-cable loop referred to above from New York to Philadelphia and back, using a picture-signal band width of 2.75 megacycles. The signals were generated by a special film scanner and transmitted on the basis of 441 lines and 30 frames interlaced. The equipment and the transmissions over the coaxial cable were witnessed by representatives of the Federal Communications Commission, members of the National Television Systems Committee panels, and several other groups interested in television. The Republican National Convention was relayed from Philadelphia to New York, via coaxial cable and several miles of special telephone cable, where it was broadcast from the television transmitter on the Empire State Building.

Around New York, increasing use was made of specially selected and equalized telephone circuits, for conveying outside-of-studio events to the station. To date, it seems quite practical to transmit pictures of acceptable quality, especially of current special events of great interest, over telephone circuits up to one mile in length.

Developments in large-picture technique continued during the year 1940. Picture tubes up to 20 inches in diameter were sold, while good projected pictures up to $4\frac{1}{2}$ by 6 feet were demonstrated. An interesting method of projecting large television pictures was described in which the screen of the picture tube has its transparency changed by the scanning electron beam. The picture-tube screen is located in the path of the light from a powerful source, so that the differences in transparency are projected on the final viewing screen to give the picture.

Active development in the field of luminescent materials and television optics continued and is likely to continue for many years. The same may be said of transmitter- and receiver-circuit developments.

Impressive color-television demonstrations were made during the year. One system presented was of the three-color type. Rapidly rotating optical color filters were employed at both transmitter and receiver. Pictures having 343 lines were used with a repetition rate of 40 per second for the colored pictures. The demonstrations gave impetus to renewed effort in this field. An interesting side light on this development is the fact that the differences in color temperature between studio and outdoor lighting as well as incorrect exposures in color films can be corrected for by the use of an electrical color monitor at the transmitter which permits independent adjustments of gain for each of the three basic color components.

Articles in German magazines would seem to indicate that television in Germany is still being carried on to a certain extent in spite of the war. The new German standard television receiver has not yet been made generally available to the public but apparently

* Decimal classification: R583.

† As of January 1, 1941, the Federal Communications Commission had issued licenses to 19 television broadcast stations and construction permits to 27 additional stations for this service.

broadcast transmissions have taken place from the central studio in Berlin and programs have also been transmitted over special wire networks to large-screen projection equipment installed in several theaters in and around Berlin. In camera-tube technique the Germans seem to have concentrated particularly on further development of the iconoscope. An article on the general development of studio technique has recently appeared in *Postarchiv* and in the same magazine another article describes the latest development in camera-tube technique. These two and other of the more-important German articles are given in the accompanying bibliography.

Test Equipment

A phase curve tracer was developed which shows directly on the screen of a cathode-ray tube, the curve of phase displacement versus frequency from 0.1 to 5.0 megacycles. This promises to be a useful instrument.

Amateur Operations

Two-inch camera and picture tubes for amateurs appeared on the market during the year and simple television receiver circuits were published. With this impetus, amateur transmissions in television began to increase. Most of this work is done in the $2\frac{1}{2}$-meter band. If history repeats itself, many important developments may be expected from these experiments.

Standards and System Analyses

The National Television Systems Committee was organized under the auspices of the Radio Manufacturers Association with the co-operation of the Federal Communications Commission. Under its supervision, nine panels of technical experts were appointed to study different aspects of television and to recommend new standards for the television industry. The work of this committee and its panels, comprising about 160 members in all, has included an intensive investigation of practically all the fundamentals and standards of television.

Several other matters were also the subject of published papers. Among these are the effect of number of lines, frame frequency, and vestigial-sideband transmission on picture quality. The effect of contrast range and gradation was likewise investigated. Also, possible uses of frequency modulation for sound or picture signals or both have been considered.

Applications

Finally, a word must be said concerning the application of television technique to other scientific fields. The electron microscope, which is a by-product of television development, has been used and improved for several years. Its resolving power far exceeds that of an optical microscope, making possible useful magnifications greater than 25,000 diameters. During the past year, this instrument has continued to make its place as a standard tool in chemical and biological research.

Television has even shown its value in astronomy. Among the most interesting of the phenomena observed during a total eclipse of the sun are the solar prominences. These resemble gigantic flames shooting outward thousands of miles from the sun's surface. During normal, noneclipse conditions, these prominences are completely submerged in a bright, steady background of sunlight and cannot be observed. However, when the sun is televised, it is possible to remove, by electrical means, the steady background of bright sunlight. When this is done the solar prominences can be seen and studied without the necessity of waiting for a total eclipse. The television apparatus used for these studies is called a "Coronaviser."

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PART VI—FACSIMILE*

Broadcast Facsimile Operation—Point-to-Point Facsimile Operation by Wire

Broadcast Facsimile Operation

Broadcast facsimile in the United States is still being conducted on substantially an experimental basis. The use of frequency modulation, both radio and subcarrier, has received considerable attention. Reproduction speeds at present range from 16 to 24 square inches per minute at 100 lines per inch definition. Dry and premoistened electrochemical recording methods are finding increasing favor.

Multiplexing of sound and facsimile services on a frequency-modulated radio-frequency carrier was the subject of investigation by a number of separate groups. The rules of the Federal Communications Commission provide† that the Commission may grant authority to such a station for multiplex transmission of facsimile and aural broadcast programs provided the facsimile transmission is incidental to the aural broadcast and does not either reduce the quality of or the frequency swing required for the transmission of the aural program. Tests have been made which indicate that satisfactory wide-range sound and facsimile broadcast services can be operated simultaneously within the 75-kilocycle band prescribed by the Federal Communications Commission for the aural program.

Point-to-Point Facsimile Operation By Wire

The newsphoto services have devoted major attention to general refinement of the apparatus and overall systems employed for transmission over telephone message toll circuits. The use of recording and playback facilities, not requiring rescanning, has been reduced to practice. Quartz-crystal-controlled frequency standards are now being used by one of the newsphoto services, replacing tuning-fork frequency standards.

Facsimile in wire-telegraph message service has continued to expand over that reported for 1939. Original facilities at New York and Chicago have been augmented by new installations and the San Francisco and Atlanta areas have been opened up for this type of wire-telegraph service. Transmitter-recorders are being used for direct patron-to-patron connection and for patron-to-central-office and vice versa. Special transmitters are installed in public centers, such as hotel lobbies, agencies, office buildings, bus depots, etc., and provide a convenient, rapid, and economical method of collecting and delivering telegrams to a central office. Automatic telegraph apparatus of the facsimile type, designed for heavily loaded circuits between branch and central offices, are in use within the areas reported. Recording units of the percussion type are in use on a number of trunk-line circuits. This type of recording, as distinguished from the chemical type, is employed where a large number of duplicate copies are desired.

* Decimal classification: R581.
† Rules governing standard and high-frequency broadcast stations. Sub-part B, rules governing high-frequency broadcast stations, Section 3.228, Federal Communications Commission No. 41741.

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PART VII—ELECTROACOUSTICS*

Loud Speakers—Microphones—Room Acoustics—Speech and Hearing—Electronic Musical Instruments—Measuring Apparatus and Techniques—Electromechanical Devices

Loud Speakers

For the most part, loud-speaker designs followed previous practices. Some refinements in moving-coil direct-radiator and horn-type loud speakers were made and these types continued their domination of radio applications.

A maximum in the power radiated from the back of

* Decimal classification: 621.385.97.

a conventional open-back cabinet occurs when the acoustic circuit looking back into the cabinet from the diaphragm is antiresonant. The use of a blocked-branch acoustic transmission line to minimize this maximum was reported.

A bellowslike diaphragm-supporting edge, having low and relatively constant stiffness, was used in a small-diaphragm moving-coil loud speaker intended for use in small enclosures.

An extension to the theory of an acoustic transmission line with finite boundary impedance to the special case of an exponential line of infinite length with finite boundary resistance was reported. This theory, when applied to the exponential horn of infinite length, indicates that even the low sound absorption of wood cannot be neglected when computing the power-transmission loss in horns of small cross section and low rates of flare. Experimentally determined values of the transmission loss of some wooden horns were used to compute the absorption coefficient of white pine and fir.

Improvements in the available energy and uniformity of the commonly used aluminum-nickel-iron family of permanent-magnet steels were effected during the year by the addition of other elements, notably copper, and by improved heat treating. At least one new permanent-magnet alloy was added to the rapidly growing list.

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Microphones

A microphone providing a total of six directivity patterns was commercially introduced. The microphone is of the two-element type and the directional characteristic is shifted by varying the relative contributions of the pressure and velocity elements. In addition to the previously available circular, cosine, and cardioid patterns, three additional quasi-cardioidal patterns are provided, thus making possible the placement of the null axis at 90, 110, 130, 150, or 180 degrees. As reported by other investigators, one of these quasi-cardioidal forms is optimum if the criterion is a maximum ratio of pickup in the desired direction to total pickup from all directions, while retaining the unilateral pickup characteristic of the simple cardioid. This pattern has been termed the "hypercardioid."

Analysis of the design factors and operating characteristics of directional microphones continued. A simple graphical method of determining the random efficiency of a microphone from measured directional characteristics was devised.

A piezoelectric microphone employing a disk of tourmaline, directly actuated by sound waves, was designed for absolute sound-pressure measurements. The generated voltage is the resultant of the piezoelectric and pyroelectric effects. A formula relating these effects to frequency was derived, enabling the computation of the response of the crystal over the audio-frequency range after certain static or low-frequency constants have been determined.

There was further study of the performance characteristics of the laryngophone (throat microphone), which is actuated by mechanical vibration of the throat walls rather than by sound waves propagated in air. The device is most useful under conditions of high ambient noise level such as prevail in military aircraft, where it has the additional advantage of freeing the pilot's hands. The crystal and carbon laryngophones, gave somewhat lower intelligibility than "ordinary" telephone apparatus when the surrounding noise was not abnormal; they gave substantially greater intelligibility under severe noise conditions.

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Room Acoustics

With the impedance concept gaining much more prominence, progress in room acoustics was mainly concerned with the application of this concept to a constantly increasing group of problems. In particular, the transmission of sound through panels and along ducts and exponential horns with absorbing boundaries was theoretically treated, with many results awaiting complete experimental confirmation.

The development of a precision curve-width method of measuring normal acoustic impedance enabled the comparison of the impedance of common absorbents. Results showed a not inappreciable and usually negative reactance component, which is being used in closer theoretical approximations, especially in a revival and modernization of Rayeigh's picture of internal absorption.

Exact analysis of diffraction by an absorbing strip was modified to correspond to practical usage and was also included in a complete solution for the behavior of a rectangular room with nonuniform application of absorbing material.

A reverberant acoustic milieu in the region of a concert artist has been simulated by immersing him in a

localized sound field supplying him an appropriately delayed, low-intensity reproduction of his rendition. The desirability of this condition is in accord with the experience of studio designers who must satisfy both artist and audience.

Current investigations along classical lines emphasized the use of reverberation time as an acoustical variable. An interesting application to the calculation of feedback conditions has been reported, and fluctuations in recorded decay curves have been smoothed by a multimicrophone method.

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Speech and Hearing

Studies on speech and hearing continued. A report on the transient characteristics of speech indicates that the rise to full amplitude may often occur in 0.05 second or less. Further light was thrown on the question of what part the vocal cords play during phonation, by the development of a technique for obtaining high-speed motion pictures of these parts. Light is concentrated on the vocal cords by a small laryngeal mirror; the same mirror reflects light back from the cords to the camera. These pictures reveal that the vocal cords do not move as units, particularly for the lower-pitched sounds.

(1) J. C. Steinberg, H. C. Montgomery, and M. B. Gardner, "Results of the World's Fair hearing tests," *Jour. Acous. Soc. Amer.*, vol. 12, pp. 291–301; October, 1940.
(2) J. F. Schouten, "The residue, a new component in subjective sound analysis," *Proc. Koninjklijke Nederlandsche Academie van Wettenscahppen*, vol. 43, p. 356; 1940; *Sci. Abst.*, sec. A, vol. 43, no. 510, pp. 489–490; June, 1940. (Abstract only.)
(3) R. O. Drew and E. W. Kellogg, "Starting characteristics of speech sounds," *Jour. Soc. Mot. Pic. Eng.*, vol. 34, pp. 43–58; January, 1940.
(4) D. W. Farnsworth, "High-speed motion pictures of the human vocal cords," *Bell. Lab Rec.*, vol. 18, pp. 203–208; March, 1940.

Electronic Musical Instruments

A solo electronic musical instrument with a six-octave range which plays a single note was announced. Frequencies in the upper octave are provided by an oscillator, the frequency of which is controlled by varying the tuning capacitance. Lower frequencies are provided by a cascade frequency divider which halves the frequency in each successive stage. The three-octave keyboard is normally attached to the keyboard of a piano. The additional three octaves are obtained by shifting the frequency the keyboard selects by an appropriate factor.

Measuring Apparatus and Techniques

Further measurements of shock waves were reported. Summaries of methods for measuring density and pressure amplitude were given including the measurement of the velocity of air flow at the wave surface by means of a filament microphone. Apparatus was also described for the dynamic calibration of crystal or other pressure gauges. The force obtained by dropping weights on a piston is transmitted through glycerine to the gauge.

An acoustic wattmeter was reported with which pressure and velocity may be measured separately as well as the energy flow represented by their product. Measurements of energy density and flow taken with this type of instrument enable the computation of the normal acoustic impedance and absorption coefficient of a surface.

(1) L. Thompson, "Shock waves in air and characteristics of instruments for their measurement," *Jour. Acous. Soc. Amer.*, vol. 12, pp. 198–204; July, 1940.
(2) C. W. Clapp and F. A. Firestone, "The acoustic wattmeter, an instrument for measuring sound energy flow." *Jour. Acous. Soc. Amer.*, vol. 12, p. 473; January, 1941. (Abstract only.)

Electromechanical Devices

Home recording, actively offered by one receiver manufacturer during 1939, was generally offered by radio-receiver manufacturers in 1940. Recording at optional speeds of $33\frac{1}{3}$ and 78 revolutions per minute was made available in home equipment. Parts manufacturers brought out improved recording devices for the home market including combination record-changer and recorder mechanisms and an inexpensive moving-armature-type magnetic cutter. Additional manufacturers entered the instantaneous-recording disk field, producing a considerable variety of paper and metal-base disks.

Further attention was given to a recording mechanism in which the linear velocity of the groove relative to the stylus is maintained constant throughout the

record. This eliminates the progressive loss in recorded high-frequency components which characterizes the common constant-angular-velocity method.

The trend in phonograph pickups toward lowered dynamical stylus-point impedance continued because of increasing emphasis on lengthening record life, reducing surface noise, and improving the reproduced response-frequency characteristic.

A phonograph pickup with a low-impedance moving system, which accomplishes transduction photoelectrically, was commercially introduced. The jeweled stylus is connected to a small mirror, which when actuated by the stylus, controls the light falling on a photoelectric cell. The exciter-lamp-filament current is supplied by a radio-frequency oscillator.

A new method of quieting the direct sound caused by the interaction of the phonograph stylus and groove and radiated from the record and pickup arm was applied to commercial phonograph cabinets. An acoustic network is formed by grooved molding, operating in conjunction with clearance space under the lid. The method provides high-frequency attenuation approaching that obtained when the cabinet lid is glued down.

(1) E. E. Griffin, "Embossing at constant groove speed—a new recording technique," *Electronics*, vol. 13, pp. 26–27, 62–64; July, 1940.
(2) "Photo-electric pickup," *Communications*, vol. 20, pp. 13–14; July, 1940.
(3) H. F. Olson, "Tone guard," *Jour. Acous. Soc. Amer.*, vol. 12, pp. 374–377; January, 1941.
(4) "Cabinet design cuts off record noise," *Radio Today*, vol. 6, pp. 42–43; October, 1940.

A number of developments previously reported had their first publication during 1940.

(1) "Artificial reverberation for electronic organ music," *Electronics*, vol. 13, pp. 42, 44; March, 1940.
(2) Benjamin Olney, "The coaxial loudspeaker," *Electronics*, vol. 13, pp. 32–35, 106–108; April, 1940.
(3) H. F. Hopkins, "An improved loud-speaking telephone," *Bell Lab. Rec.*, vol. 18, No. 8, pp. 251–254; April, 1940.

PART VIII—RADIO WAVE PROPAGATION*

The growing importance of the ultra-high-frequencies (over 30 megacycles) was emphasized by increasing activity in the determination of the facts of wave propagation at those frequencies. An extended study at 75 and 150 megacycles demonstrated that there is considerable variation of received intensity (as much as two to one) even over optical paths. It was indicated that the fading is due to changes in atmospheric refraction and to changes in tropospheric reflection at air-mass boundaries.

Variations of average received field, as affected by vegetation and buildings and other objects, were determined. Buildings and irregular terrain caused the average field intensity to vary by ten to one.

Extensive data on ultra-high-frequency wave propagation were presented at the hearings before the Federal Communications Commission on television (January 15, 1940) and on aural broadcasting at frequencies above 25 megacycles (March 18, 1940). At the latter hearing there was presented a comprehensive theoretical treatment of the effect of the troposphere on radio transmission at these frequencies.

The frequency range of measuring apparatus was extended upward. Equipment became generally available for field-intensity measurement up to 125 megacycles and for voltage measurement up to 300 megacycles.

There was some progress in the development of instruments for measuring noise intensity. Absolute measurements were made showing the noise intensity of automobile ignition over a wide range of frequencies; it was found that the magnitude is considerable even up to 450 megacycles.

Knowledge of ionospheric transmission was steadily augmented. Predictions of maximum usable frequencies for various distances and times of day were published each month in the PROCEEDINGS of the I.R.E., and were found to be of sufficient reliability to aid in planning transmitter frequency schedules. The correlation of radio-transmission conditions with geomagnetism was facilitated by the inauguration of more convenient reports and indexes of geomagnetic activity. The reception of echoes was reported from distant regions of the ionosphere where there is a marked curvature of one of the layers. It was discovered that sudden ionospheric disturbances occasionally affect broadcast frequencies, sometimes decreasing and sometimes increasing received intensities.

The steerable antenna system, to reduce fading caused by transmission over multiple paths, came into commercial use. It was found under certain conditions to improve quality of reception by increasing the vertical directivity of the receiving antenna so as to favor the waves arriving at one angle to the exclusion of others.

(1) C. R. Englund, A. B. Crawford, and W. M. Mumford, "Ultra-short-wave transmission over a 39-mile 'optical' path," PROC. I.R.E., vol. 28, pp. 360–369; August, 1940.
(2) B. Trevor, "Ultra-high-frequency propagation through woods and underbrush," *RCA Rev.*, vol. 5, pp. 97–100; July, 1940.
(3) G. S. Wickizer, "Field-strength survey, 54.75 megacycles from Empire State Building," PROC. I.R.E., vol. 28, pp. 291–296; July, 1940.
(4) K. A. Norton, "A theory of tropospheric wave propagation," Federal Communications Commission hearing of March 18, 1940, FCC No. 40030.
(5) R. W. George, "Field-strength measuring equipment at 500 megacycles," *RCA Rev.*, vol. 5, pp. 69–76; July, 1940.
(6) C. V. Aggers, D. E. Foster, and C. S. Young, "Instruments and methods of measuring radio noise" (Based on recommendations of Joint Co-ordination Committee on Radio Reception of EEI, NEMA, and RMA), *Trans. A.I.E.E.* (*Elec. Eng.*, March., 1940), vol. 59, pp. 178–182; 1940.
(7) R. W. George, "Field strength of motorcar ignition between 40 and 450 megacycles," PROC. I.R.E., vol. 28, pp. 409–412; September, 1940.
(8) J. Bartels, N. Heck, and H. P. Johnston, "The three-hour-range index measuring geomagnetic activity," *Terr. Mag.*, vol. 44, pp. 411–454; December, 1939.
(9) J. A. Pierce and H. R. Mimno, "Reception of radio echoes from distant ionospheric irregularities," *Phys. Rev.*, vol. 57, pp. 95–105; January 15, 1940.
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* Decimal classification: R113.7.

Some Notes on Linear and Grid-Modulated Radio-Frequency Amplifiers*

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Summary—*It is shown that the regulation of the radio-frequency exciter of a grid-modulated amplifier can be greatly improved by connecting across the exciter a diode biased so that at the peak of the modulation cycle the diode is just beginning to draw current. The exciter can then be designed on the basis of the peak exciting power required, and without regard to regulation. The method is also extended to include compensation for the flattening of the positive peaks of the audio-frequency modulating voltage. This is done by using a triode limiting tube that is so arranged that grid current in the modulating stage causes the load on the exciter to be reduced.*

A method of applying feedback to an isolated linear amplifier stage (or stages) is described. This consists in rectifying samples of the input and output radio-frequency voltages, balancing the resulting audio frequencies in the rectifier outputs against each other, and modulating the difference that results upon the system in such a way as to tend to correct for the distortion causing the unbalance. This system can be applied to linear amplifiers that are added to a transmitter subsequent to installation without disturbing in any way the negative feedback system of the existing transmitter.

LINEAR stages of radio-frequency amplification are frequently used in modern radio broadcast equipment. Their use results in a particularly flexible unit-construction transmitter design, allowing the addition of higher-power stages subsequent to the original installation with a minimum of trouble and expense, and with the possibility of maintaining high over-all efficiency.[1] Likewise, grid modulation has found use in radiotelephone equipment because of the low modulating power required, and has become especially attractive with the development of high-efficiency systems.[2]

This paper describes several ways in which the performance of such systems may be improved. These include a new means of obtaining the advantages of negative feedback in a linear amplifier. Also methods are described whereby in a grid-modulated amplifier operated so that the grid draws current at the peak of the modulation cycle, the power capacity of the sources of exciting and modulating voltages can be kept at a minimum.

Exciter Systems for a Grid-Modulated Radio-Frequency Amplifier

One of the important limitations involved in minimizing distortion in the grid-modulated radio-frequency amplifier is the problem of maintaining good regulation in the radio-frequency driving voltage. In order to obtain good efficiency and appreciable power output in proportion to the tube size, it is necessary that the grid be driven positive at the crest of the modulation cycle, just as in the case of the ordinary class C amplifier. This causes grid current to flow at and near the crest of the modulation cycle and this grid current represents an added load on the radio-frequency driving voltage. This reduces the radio-frequency exciting voltage, thus flattening the positive peaks of modulation with a consequent introduction of distortion. The voltage and current relations that exist in a typical case are shown in Fig. 1, the left-hand column representing the condition of perfect driver regulation, and the right-hand column representing the practical condition of imperfect driver regulation.

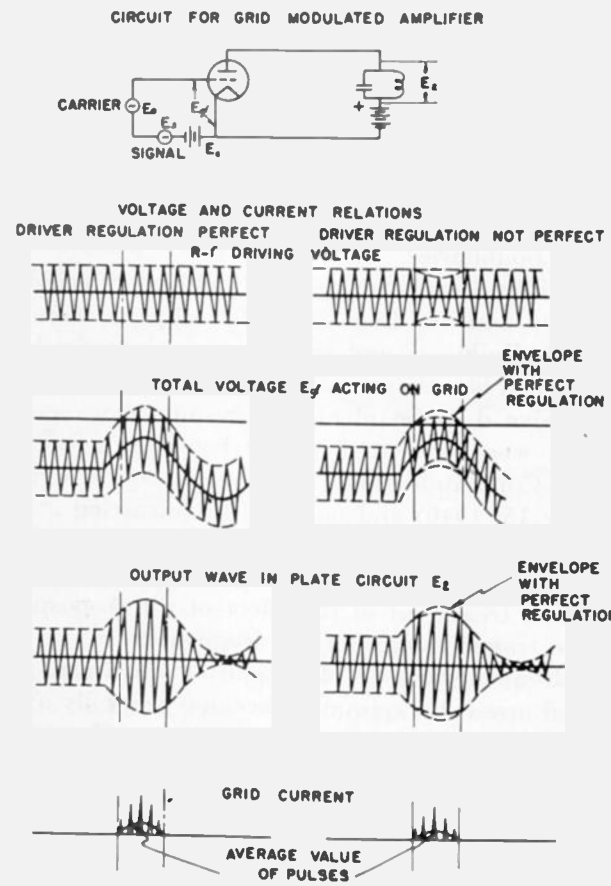

Fig. 1—Voltage and current relations in a grid-modulated amplifier showing the effect of imperfect driver regulation.

If by some means the load on the driver could be maintained substantially constant while still allowing the grid-modulated amplifier to draw grid current, the regulation of the driver would no longer be of importance, and modulation free from this kind of distortion

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† Stanford University, Calif.

‡ Formerly, Stanford University, Calif.; now, Litton Engineering Laboratories, Redwood City, Calif.

[1] W. N. Doherty, "A new high-efficiency power amplifier for modulated waves," Proc. I.R.E., vol. 24, pp. 1163–1182; September, 1936.

[2] F. E. Terman, and John R. Woodyard, "A high-efficiency grid-modulated amplifier, Proc. I.R.E., vol. 26, pp. 929–945; August, 1938.

could be obtained. To obtain such a result, it has been customary to provide a driver capable of generating much more power than required to excite the grid-modulated amplifier, and then to load the driver with a resistance so that the added load due to the grid current at the modulation peaks was a small part of the total load on the driver.

It is proposed here to provide a driver that is barely able to supply the peak driving power required by the grid-modulated amplifier, and then to maintain the load on the exciter substantially constant by the use of a limiter as illustrated in Fig. 2. Here a diode is shunted across the radio-frequency driver and provided with a positive bias on the cathode. This bias is adjusted so that at the peak of the modulation cycle when the grid of the modulated tube is most positive, the diode draws little if any current. That is to say, the bias is adjusted to a value approximately equal to the peak amplitude of the radio-frequency driving voltage desired for the grid-modulated amplifier. At the peaks

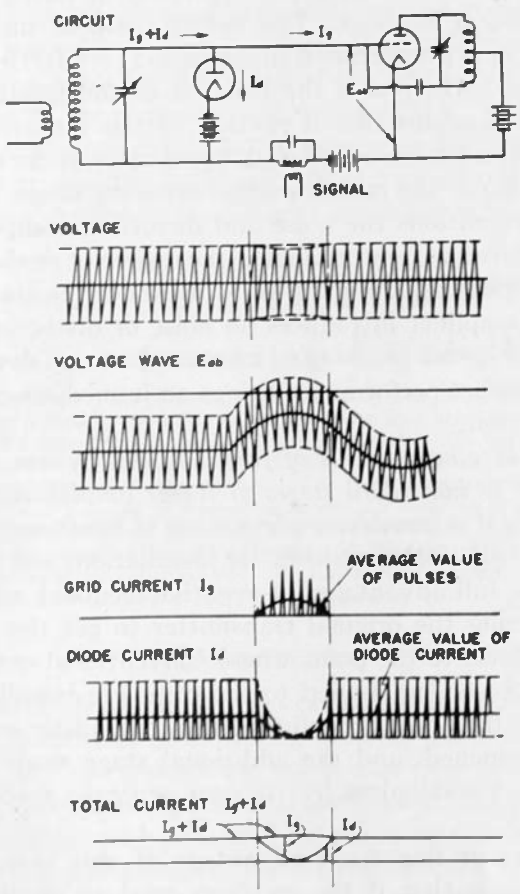


Fig. 2—Voltage and current relations in a grid-modulated amplifier with a diode regulator tube added to control driver regulation.

of the modulation cycle the diode has little or no effect on the driver. At any other part of the modulation cycle, the radio-frequency exciting voltage tends to rise because of the reduced load presented by the grid of the modulated tube. The presence of the biased diode, however, greatly minimizes this tendency for the exciting voltage to increase, since even a small increase in voltage causes the diode to draw appreciable current at the peaks of each radio-frequency cycle, thereby placing a load upon the driver that increases rapidly as the driver voltage increases. The result is to improve greatly the constancy of the exciting voltage.

The detailed mechanism involved is illustrated in the lower part of Fig. 2. If the plate resistance of the

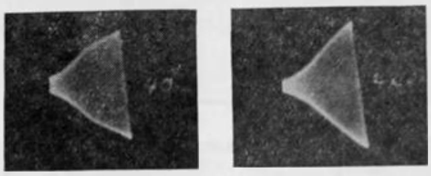


Fig. 3—Oscillograms showing the improvement in operation of a grid-modulated radio-frequency amplifier with the addition of a diode to control driver regulation. The trapezoid to the left is for a conventional arrangement, and has reduced slope at large amplitude because of poor regulation of the exciting voltage. In contrast, the trapezoid to the right has substantially straight sides at large amplitude as the result of the action of a diode regulator.

diode is low, as is the case with rectifier tubes now available, the peak radio-frequency exciting voltage need rise only a small amount above the diode bias to cause a large current to flow through the diode, with correspondingly large losses being placed upon the exciter. The result is that the regulation can be made extremely good in practice, and it becomes entirely feasible to design the driver stage to provide only sufficient power to operate the grid-modulated tube at the peak of the modulation cycle, and to ignore entirely the matter of regulation by allowing the diode regulator tube to take care of this requirement.

The oscillograms of Fig. 3 show the very considerable improvement in operation of a laboratory setup obtained through the use of such a diode regulator.

This same method of approach can be extended to correct for the flattening of the positive peaks of the audio-frequency modulating voltage caused by grid current, as shown in Fig. 4. Here a triode voltage-regulating tube A performs a function similar to that

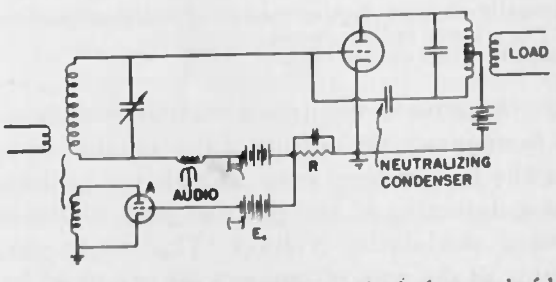
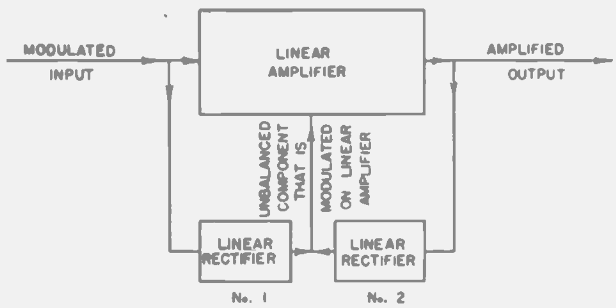


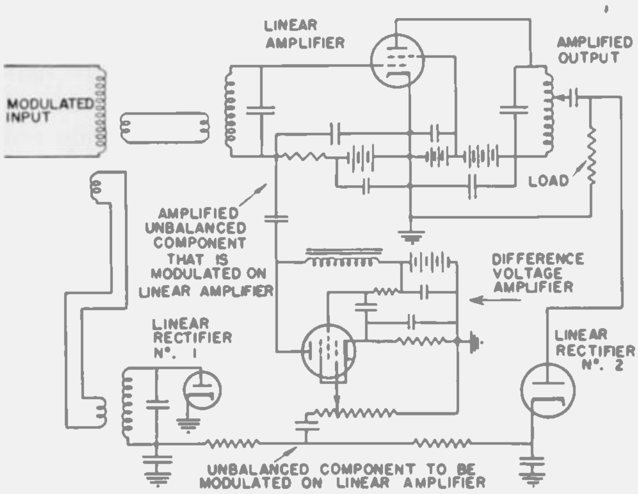
Fig. 4—Schematic circuit showing method of control of both driver regulation and modulator regulation.

of the diode tube in Fig. 2 except that the bias for the regulator tube is obtained from a combination of battery E_c and grid leak R. The adjustment is such that during the unmodulated period when the grid current is zero then the bias E_c is such that the plate of tube A is driven sufficiently positive to cause this tube to absorb a moderate amount of power. If it were not for the grid-leak resistance R the system would then func-

tion as a regulating device exactly as does the diode tube in Fig. 1, and thereby maintain good regulation of the radio-frequency exciting voltage. However, because of the grid-leak resistance an additional bias is applied to A whenever the modulation is sufficient to cause the grid of the modulated amplifier to become positive. This additional bias reduces the loading that A places on the exciting voltage and allows the exciting

(a)





(b)

Fig. 5

(a) Schematic diagram showing method of applying balanced feedback to a linear radio-frequency amplifier.
(b) Practical circuit used in tests.

voltage to increase when grid current is drawn. This tends to maintain the output of the modulated amplifier at the proper level even though the grid current causes a flattening of the positive peak of the audio-frequency modulating voltage. That is to say, the flattening of the audio-frequency wave caused by grid current is compensated for by using this same current to increase the radio-frequency exciting voltage.

Distortion Reduction in Linear Amplifiers

All of the benefits given by conventional negative feedback can be obtained with an isolated linear amplifier by the method shown in Fig. 5. Here a sample of the modulation envelope is obtained from the output by means of a linear detector, and balanced against a similar sample obtained from the input. The resulting difference voltage then consists of noise and distortion components that appear in the output but not in the input, that is, the difference voltage contains a sample of the distortion products of the amplifier. This difference voltage is then "remodulated" on the radio-frequency wave with such polarity as to tend to cancel the original noise and distortion components appearing in the output wave. The introduction of this corrective balance voltage as remodulation is the radio-frequency equivalent of a balanced-feedback amplifier.[3]

The performance of such an amplifier with feedback introduced is the same as an ordinary linear amplifier except that the noise and distortion introduced by the amplifier are reduced by a factor $1/(1-Ak\beta)$, where A is the voltage gain of the linear amplifier, β is the fraction of the output returned to be balanced against a portion of the input, and k is the transmission factor of the network subsequent to the balance point. Since rectification is included in the feedback path, the values of A, k, and β are expressed in terms of the modulation envelope. The voltage gain of the linear amplifier is proportional to the ratio $(1-ka)/(1-Ak\beta)$ however, where a is the fraction of the input to be balanced against the β portion of the output. Ordinarily this ratio is adjusted to equal unity under normal operation in the middle audio-frequency range. Under these conditions the noise and distortion components are reduced without changing the degree of modulation or otherwise affecting the performance of the amplifier. If the amplifier introduces no noise or distortion, the *feedback circuit has no effect whatever*, but any deviation from perfect performance brings an immediate corrective action.

Unlike other systems of feedback, this system can be applied to individual stages of linear amplification. In this way it is possible to add a stage of linear amplification to an existing transmitter installation, and to obtain the full advantages of negative feedback without redesigning the original transmitter to get the phase shifts down to the point where conventional envelope feedback can be applied to the complete installation. That is, the original portion of the transmitter may be left untouched, and the additional stage made sufficiently distortionless by its own separate system of feedback.

A further important advantage of this system of feedback is that if the rectifiers used to obtain the modulation envelopes of the input and output are identical in characteristics, *any distortion of these envelopes caused by the rectifiers cancels at the balance point and does not appear in the difference voltage to be introduced into the radio-frequency amplifier.*

Since only a single stage need be involved, phase shifts are so low that the difference voltage may be amplified before reinsertion into the amplifier in order

[3] E. L. Ginzton, "Balanced feed-back amplifiers," Proc. I.R.E., vol. 26, pp. 1367–1379; November, 1938.

to increase the corrective action. The schematic circuit of Fig. 5 shows such an amplifier and the experimental oscillograms of Fig. 6 show the great improvement in performance of a linear radio-frequency amplifier that was obtained using a circuit similar to that of Fig. 5.

This method of balanced feedback, with modifications, may also be applied to other types of radio-frequency amplifiers such as the grid-modulated amplifier discussed in the preceeding section of this paper, but it is particularly effective when applied to a linear amplifier, and avoids the difficulties of many other means of minimizing distortion.

Conclusions

The results of this paper may be summarized as follows: (1) The regulation of the exciter of a grid-modulated amplifier can be greatly improved by means of a limiting system, which reduces the required power capacity of the exciter to a value little if any greater than that required to supply the peak driving power; (2) this method can be extended to provide an overcompensation so that the exciting voltage of the grid-modulated amplifier actually rises instead of decreases when the grid current is drawn, thus making is possible to correct for flattening of the audio modulating voltage at the peaks of modulation; (3) a "remodulation" system of distortion correction is described that makes it possible to obtain all of the advantages of negative feedback in a linear amplifier added to a transmitter without the necessity of making any changes in the feedback system of the original transmitter. These possibilities offer opportunities for increased economy, flexibility, and simplification in the design of radiotelephone transmitters.

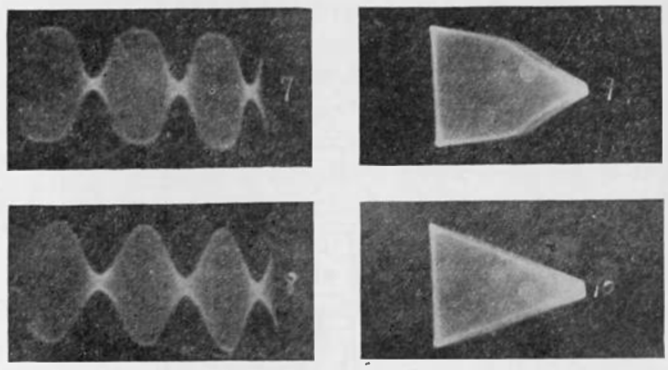


Fig. 6—Oscillograms showing the improvement in operation of a linear amplifier with the addition of balanced feedback. The upper line shows modulation envelope and trapezoid pattern for a conventional linear amplifier having considerable distortion. The second line is for exactly the same conditions except that the balanced-feedback system has been connected.

Additional References

(1) H. S. Black, "Stabilized feed-back amplifiers," *Elec. Eng.* vol. 53, p. 114; January, 1934.

(2) H. Nyquist, "Regeneration theory," *Bell Sys. Tech. Jour.*, vol. 11, p. 126; January, 1932.

(3) H. A. Thompson, "A stabilized amplifier for measurement purposes," *Trans. A.I.E.E*, vol. 57, p. 379; July, 1938.

(4) P. O. Pedersen, "A distortion-free amplifier," Proc. I.R.E., vol. 28, pp. 59–66; February, 1940.

A Phase Curve Tracer for Television*

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Summary—This phase curve tracer is a system for showing on the screen of a cathode-ray tube the phase curve of any network, plotted on a linear frequency scale from 0.1 to 5 megacycles. The test frequency is changed to a fixed frequency of 50 kilocycles for phase comparison, and the phase shift is converted to a time shift. A rectangular field on the screen is scanned in vertical lines, one for each test frequency. A bright spot is produced on each line at a vertical distance proportional to the phase angle of the circuit under test. Frequency and phase co-ordinate lines are superimposed. The full scale of phase indication is adjustable in multiples of 360 degrees by switching.

Introduction

The desirability of investigating the phase characteristics of television systems has been generally recognized. However, the time involved in making the necessary measurements with usually available equipment has limited such investigations. A solution for the situation lies in the construction of a phase curve tracer which produces a curve showing the variation with frequency of the phase angle between the input and output voltages of a system.

The device to be described shows on the screen of a cathode-ray tube the phase characteristics of any circuit between 0.1 and 5 megacycles. Frequency and phase co-ordinates are also produced, giving a visual effect of a phase curve plotted on graph paper.

To trace out a phase curve on a cathode-ray tube with linear co-ordinates, a horizontal sweep is needed which produces a deflection that is a linear function of frequency, together with a phase-angle indicating device which produces a vertical displacement that is a linear function of phase angle. A device which indicates phase angles by measuring the time interval between corresponding reference points, such as intercepts or peaks of the two sine waves, can be made to fulfill the latter requirement; but it is preferable that this type of indicator operate at a constant frequency. Thus the apparatus to be described can broadly be divided as follows:

The Indicator Device, which operates at a constant frequency (50 kilocycles), like the intermediate frequency in a superheterodyne receiver, and produces a phase-angle indication on cathode-ray tube.

The Signal Generator, which generates the variable-frequency investigating signal (0.1 to 5 megacycles), and also produces two constant intermediate-frequency signals—one a reference derived from the signal input to the circuit under test, and the other a "test" signal derived from the output signal of the circuit under test.

* Decimal classification: R583. Original manuscript received by the Institute, January 8, 1941. Presented, Rochester Fall Meeting, November 11, 1940, and Sixteenth Annual Convention, New York, N. Y., January 11, 1941.

† Hazeltine Service Corporation, Little Neck, L. I., N. Y.

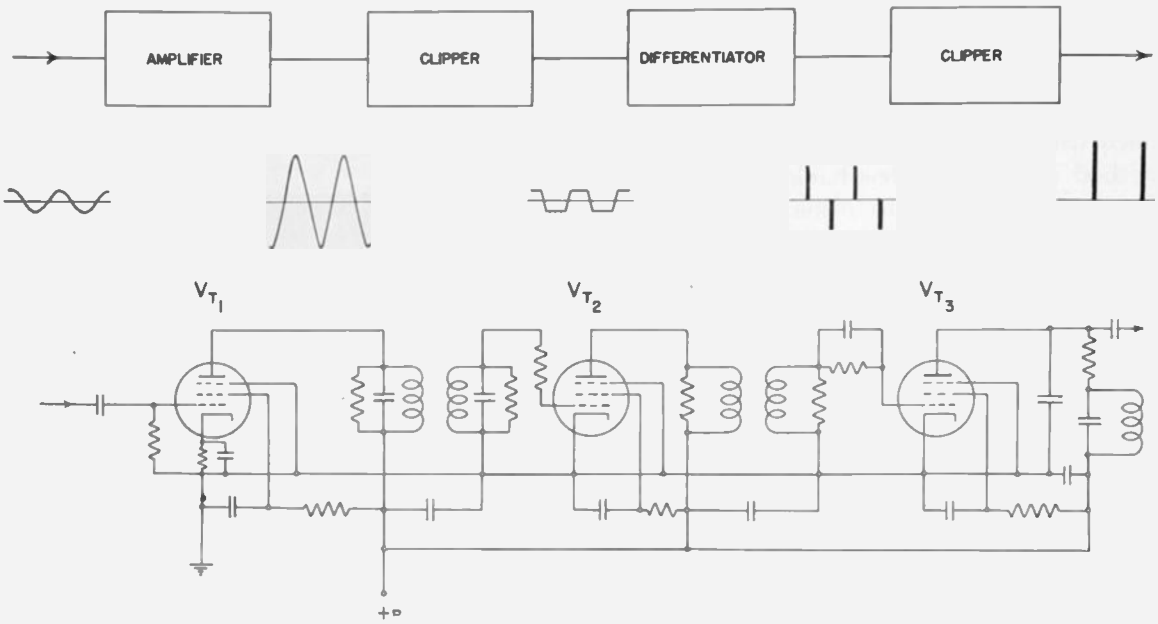


Fig. 1—Intercept pulse generator.

The Horizontal Scanning Unit, which produces a deflection on the cathode-ray tube that is a linear function of frequency.

Indicator

The indicator unit of this apparatus measures phase angles by measuring the time elapsed between corresponding intercepts of the two sine waves being tested. A linear time axis is produced vertically on the cathode-ray tube by a linear saw-tooth deflecting voltage, which is accurately synchronized with one sine wave so that its retrace occurs at the upward intercept. Then a short pulse is produced at the upward intercept of the other sine wave, and this is applied to the grid of the cathode-ray tube, thus modulating the beam. The result is a vertical trace with a spot on it; and since the deflection of the beam along the trace is a linear function of time, the distance of the spot from the starting end of the trace will be directly proportional to the phase angle between the two signals. The vertical sweep can be accurately synchronized with one sine wave by first producing a short pulse from the upward intercept of that sine wave and using this pulse to synchronize a linear saw-tooth oscillator.

These short pulses from the intercept of a sine wave which are needed for both synchronizing of the sweep and modulation of the beam can be produced by a simple wave-shaping process. For convenience, the unit producing this wave shaping will be called an "intercept pulse generator." (See Fig. 1.) First the signal is amplified so that between 100 and 200 volts, peak to peak, appears across the secondary of the transformer at $x-x$. Then this is drastically limited by grid saturation in the positive direction and plate-current cutoff in the negative direction, so that the plate current of V_{T2} is a square wave. Next, the square wave is differentiated by the mutual inductor between V_{T2} and V_{T3}, producing double pulses as shown; and finally V_{T3} is a limiter or a peak detector so that its output current contains only one polarity of pulse. The plate circuit of V_{T3} contains a low-pass filter used to maintain sufficient band width to transmit the pulse without distortion.

One factor determining the accuracy with which the phase-angle indicator can be read is the width of the pulse as a per cent of the vertical scanning cycle, since this affects the size of the spot produced by the pulse. It is desirable to have the pulse at least as narrow, or narrower, in degrees than the desired accuracy in degrees. Thus for a 2-degree accuracy, the pulse should not be more than 1/180th of a cycle. Since the duration of the pulse corresponds approximately with half the period of the highest frequency that must be transmitted for accurate reproduction of the pulse, the band width of circuits following the intercept pulse generator is determined. Accordingly in this apparatus where phase angles are measured at 50 kilocycles, and

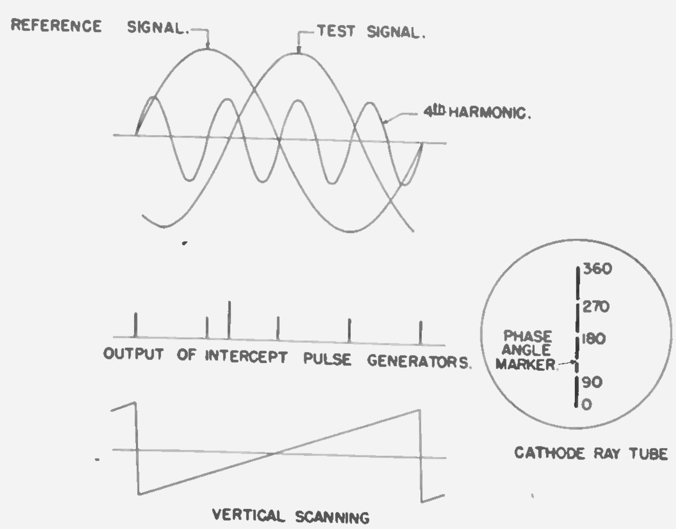


Fig. 2—Phase-angle markers.

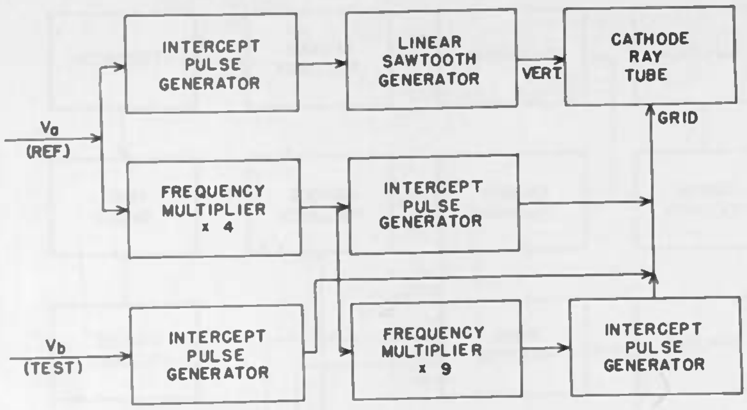


Fig. 3—Simplified block diagram of indicator.

a 2-degree accuracy of reading is desired, the circuits between the intercept pulse generator and the grid of the cathode-ray tube have a 5 megacycle band width.

By evenly dividing the length of the vertical trace, a scale can be made from which the phase angle can be read directly from the location of the spot with reference to the indexes of the scale. These phase-angle markers can be made a part of the vertical trace by further modulation of the cathode-ray beam. Thus by applying four equally spaced pulses per cycle of the vertical scan to the grid of the cathode-ray tube, four markers are produced on the trace which indicate points that are 90 degrees apart. These four equally spaced pulses per cycle can be produced by applying the reference sine wave (i.e., the one used to synchronize the linear saw-tooth sweep) to a frequency-multiplying stage, selecting the fourth harmonic, and putting this through an intercept pulse generator (see Fig. 2). Thus with 50 kilocycles as the intermediate frequency, these 90-degree markers are really short 200-kilocycle pulses.

Further phase-angle markers can be produced by further frequency multiplications. In the apparatus being described, 10-degree phase markers were also added. This means that the 200-kilocycle signal was multiplied by 9 and applied to an intercept pulse generator, giving 1.8-megacycle pulses (i.e., 36 pulses per vertical scanning cycle, or one every 10 degrees). By making the 90-degree markers of greater amplitude than the 10-degree markers, the appearance of major and minor subdivisions on a scale is produced.

Fig. 3 shows a simplified block diagram of the indicator unit as described up to now. In order to have this indicator produce an accurate measure of the phase angle between V_a and V_b, the timing of all the various pulses must be correct. Thus when V_a and V_b are in phase (or the same signal) the phase-angle indicator spot should be at the bottom of the trace, one of the 90-degree markers should coincide with the phase-angle indicator spot, and every 9th of the 10-degree markers coincide with a 90-degree marker. The practical problems involved here, along with those concerning harmonic generators and linear saw-tooth oscillators, should make it evident why this indicator unit is operated at a constant frequency, and not directly from the 0.1-to-5-megacycle signal going through the test circuit. Operation at a constant frequency also makes the problem of phasing the various pulses quite simple, since simple trimmers on various tuned circuits can be used.

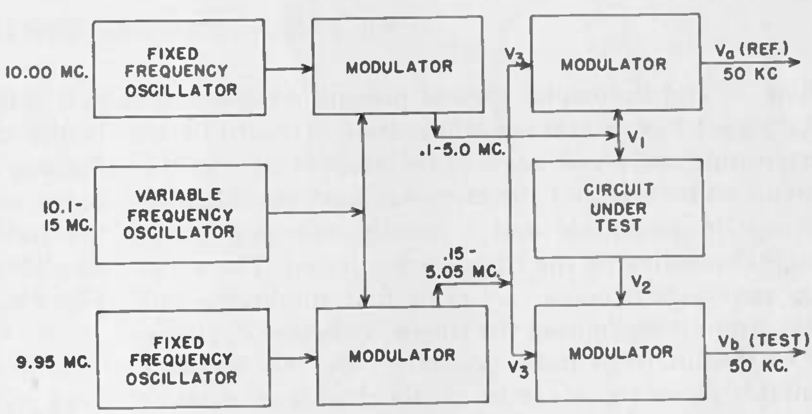


Fig. 5—Simplified block diagram of signal generator.

A practical working model of this indicator unit requires the addition of amplifiers and buffers (for freedom from interaction of controls), as well as zero sets for proper phasing of pulses. The expanded block diagram of such a unit is shown in Fig. 4.

The reference signal V_a, as produced by the signal generator, is of approximately constant amplitude for various frequencies of investigation of the test circuit, and has only a slowly changing phase angle when the investigating frequency is swept through the 0.1-to-5-megacycle range. Thus the design of circuits han-

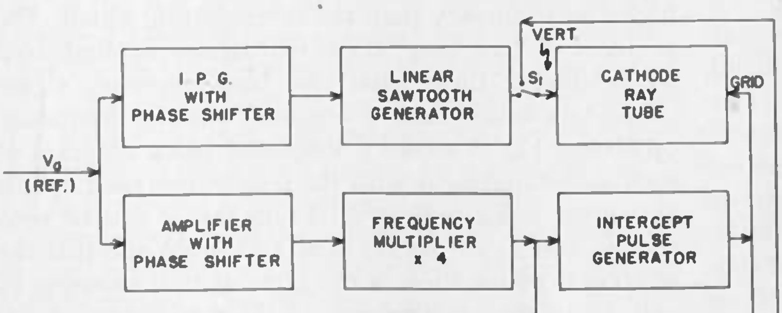


Fig. 4—Expanded block diagram of indicator.

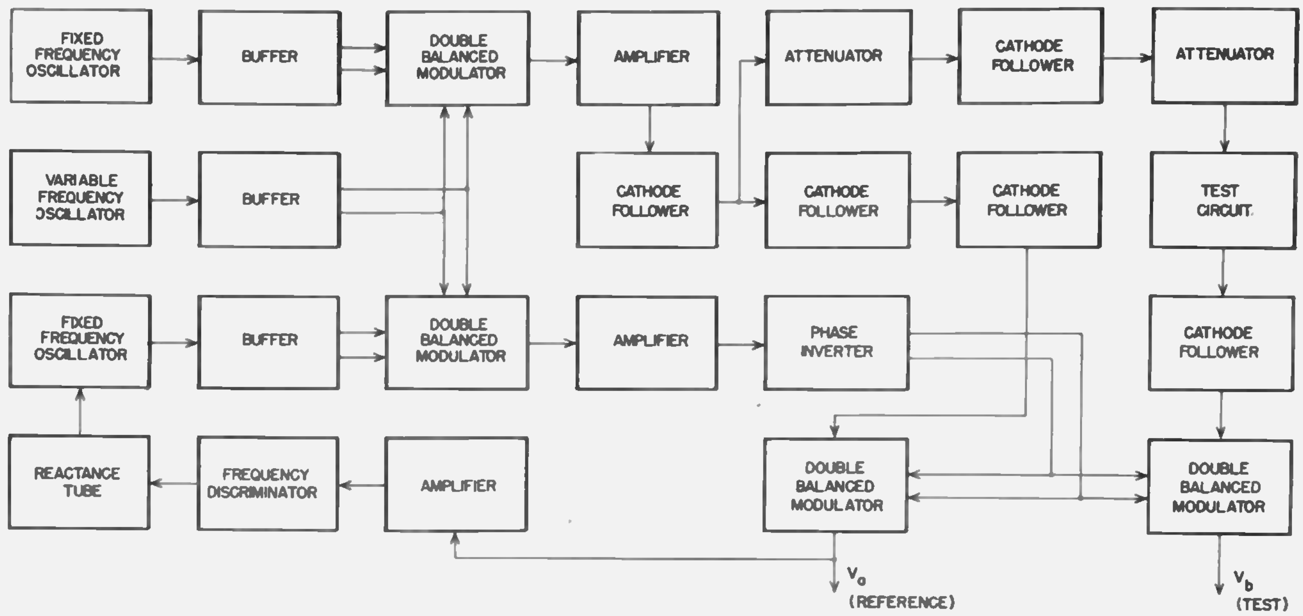


Fig. 6—Expanded block diagram of signal generator.

dling V_a and harmonics thereof present no difficulty. But signal V_b contains the information in regard to the magnitude and phase angle of the amplification of the circuit under test and therefore can have considerable change in amplitude and a rapidly changing phase angle depending on the circuit being tested. The varying amplitude is taken care of by first amplifying and then amplitude-limiting the sine wave before applying it to the intercept pulse generator. By this means a suitable phase curve can be obtained with as much as a 40-decibel variation in amplitude. However, the present amplitude-limiting system introduces some errors which are the limiting factors in accuracy of indication. A variation in amplitude of V_b over the 40-decibel range produces a ± 3-degree error in indication. There is a further error introduced due to the rate of change of amplitude of V_b, and this amounts to ± 5 degrees when sweeping through a trap circuit of reasonable Q.

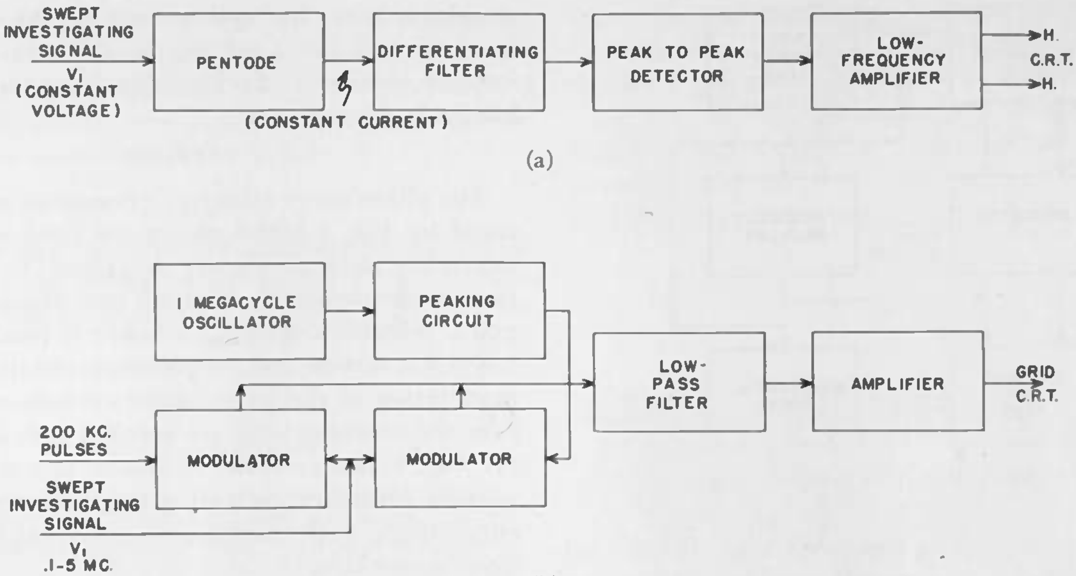
The rapidly changing phase angle of V_b requires that the channel handling this must have sufficient band width and linear phase characteristics in order not to distort this phase- and amplitude-modulated signal. The considerations involved include not only the maximum slope of the phase curve, but also the rate of frequency sweep in the test circuit. In the present apparatus, a band width of 20 kilocycles is used and this channel is consequently substantially flat from 40 to 60 kilocycles, this band width being adequate to handle signals with phase slopes of the order of those in an over-all television system and those produced by trap circuits with a reasonable Q. This figure is somewhat conservative in order to insure a linear phase shift (i.e., a constant time delay) of all essential components of the phase-amplitude-modulated signal V_b.

The phase-angle indicating device just described can also be used to indicate the amplitude response of the circuit under test. For this, the vertical deflecting circuit of the cathode-ray tube is connected to some point in the channel transmitting V_b before any amplitude limiting is produced. Then if the input to the circuit under test is constant at all frequencies, the length of the vertical trace will indicate the magnitude of the amplification of the circuit under test. Switch S_1 in Fig. 4 indicates this change.

Signal-Generator Unit

The signal-generator unit should produce a variable-frequency investigating signal (0.1 to 5 megacycles) and also two constant-frequency signals (50 kilocycles), one a reference derived from the input and the other a "test" signal derived from the output of the circuit under test. These constant-frequency signals are produced by heterodyning the investigating signal with a "beating" signal which is always 50 kilocycles higher in frequency than the investigating signal. The problem of "tracking" these two signals is eliminated by producing them from two beat-frequency signal generators which have a common variable-frequency oscillator. Fig. 5 shows a simplified block diagram of such an arrangement with the frequencies used in this apparatus being indicated.[1] From this it can be seen that V_a and V_b are always at 50 kilocycles, and that the relation between them is the same as that between V_1 and V_2, or the amplification of the test circuit. Actually with the beat signal higher in frequency than the investigating signal, the phase angle between V_b and V_a is the negative of that between V_2 and V_1. This merely means that the vertical sweep on the cathode-ray tube must be scanned from the top down, in order to make a lagging phase angle between V_2 and V_1 produce an upward motion of the spot. If signals V_1 and

[1] For a similar signal-generator unit, see M. Levy, "Methods and apparatus for measuring phase distortion," *Elec. Comm.*, vol. 18, pp. 206–228; January, 1940.



(a)

(b)

Fig. 7

(a) Horizontal sweep circuit. (b) Frequency-marker circuit.

V_2 are constant in amplitude at all frequencies, then V_a is constant and the amplitude of V_b varies directly as the magnitude of amplification of the circuit under test.

A practical working model of a generator like this requires the use of amplifier buffers, double balanced modulators, impedance transformers (i.e., cathode followers), and attenuators in various places. Fig. 6 shows an expanded block diagram of the unit constructed for this apparatus. The cathode followers before the test circuit prevent interaction between the low-impedance attenuator and the high-impedance low-pass filter of the previous amplifier, and also produce a low-impedance output so that test circuits with the usual order-of-input impedance do not affect the signal output of the generator. The cathode follower after the test circuit is in a "test probe" reflecting a high input impedance to the test circuit and a low impedance to the transmission feeding the signal back to the modulators. These impedance-transforming circuits are also simulated in the channel feeding the reference-signal modulator, so that the only essential difference between the paths traveled to produce V_a and V_b is that the latter include the circuit under test and attenuators.

Proper operation of the indicating unit requires that the constant-frequency signals be kept quite accurately at 50 kilocycles. This frequency is determined by the difference frequency of the two fixed high-frequency oscillators. When using self-excited oscillators it is desirable to have some form of stabilization, and accordingly an automatic-frequency-control circuit is used as shown so that the difference frequency is stabilized by the 50-kilocycle discriminator.

Stray coupling produces an extraneous 50-kilocycle signal in the output of the first pair of double balanced modulators, which must be prevented from getting into V_a and V_b. This is accomplished by resistance-compensated 50-kilocycle traps in the investigating-signal and beat-signal channels, and also by an accurate balance of the second pair of double balanced modulators.

Sweep Operation

To complete the phase curve tracer, it now remains to produce all those operations connected with the horizontal scanning of the curve. This includes periodic variation or sweeping of the investigating frequency, production of horizontal scanning for the cathode-ray tube, and the addition of frequency markers which, in conjunction with the phase-angle markers, complete the illusion of graph paper. Sweeping the investigating frequency is a simple problem since only one oscillator frequency is varied. In this apparatus a motor-driven condenser is used which is operated synchronously at 20 cycles, going through the 0.1-to-5-megacycle range in 1/40th of a second.

The horizontal sweep is produced by a frequency-discriminator arrangement operated directly from the investigating signal. As indicated in Fig. 7(a), a differentiating filter is used which, when supplied with a constant current, produces an output voltage whose amplitude is a linear function of frequency. Then the envelope of this amplitude-frequency-modulated signal is detected and the output is used for the horizontal sweep. This sort of arrangement has the advantage that it will produce a horizontal sweep which is a linear function of frequency regardless of the shape of the frequency-deviation-versus-time curve. Thus the motor-driven condenser used to sweep the frequency need not have specially shaped plates. The differentiating circuit accentuates any harmonic distortion that may exist, and unless a peak-to-peak detector is used, a non-

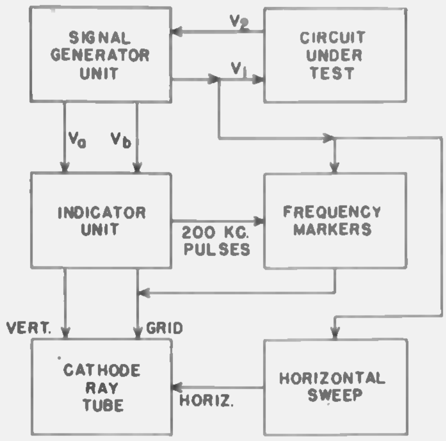


Fig. 8—Block diagram of complete system.

linearity of trace may be produced when the second-harmonic distortion component goes out of the pass band of the system.

Frequency markers are produced by beating the swept-frequency investigating signal with standard-frequency signals and brightening up several vertical scanning lines as the signals go through zero beat. The block diagram of this circuit is shown in Fig. 7(b). A 1-megacycle oscillator and its harmonics produce markers at every megacycle. Since it is desirable that all harmonics of interest be of approximately equal amplitude, a 1-megacycle pulse of short duration should be supplied to the modulator. Accordingly the plate-current pulses of the 1-megacycle oscillator are fed to a peaking filter which partially differentiates them. Markers are also added every 200 kilocycles, using the 200-kilocycle pulses which produce the 90-degree phase-angle markers as the standard. The result is again the appearance of major and minor subdivisions along the frequency axis. It will be noted that if the 1-megacycle oscillator is an accurate frequency standard, then the coincidence of the 200-kilocycle markers and the 1-megacycle markers offers a convenient means of checking the frequency in the 50-kilocycle channel.

System

The phase curve tracer as a complete system is indicated by Fig. 8 which shows the interconnection between the units previously described. It is interesting to note the similarity between this phase curve tracer and a television system. A raster is produced as in a television system and the picture is obtained by proper modulation of the beam of the cathode-ray tube, but here the scanning lines are vertical instead of horizontal. Fig. 9 shows a raster of coarse line structure with varying brilliancy indicating the formation of a phase curve along with 90-degree markers and several frequency markers.

A photograph of the phase curve tracer is shown in Fig. 10. The apparatus is built in three units which are mounted on a 19-inch relay rack, and for portability this relay rack is mounted on a dolly. The top unit contains the indicator, horizontal-sweep, and frequency-marker circuits, as well as the cathode-ray tube and its power supplies. The middle unit contains the signal-generator circuits. At the lower right of this unit can be seen the cable which feeds the 0.1-to-5-megacycle signal to the test circuit, and also the gooseneck with the probe at the end which receives the output signal of the test circuit. The bottom unit contains the power supplies to furnish filament and

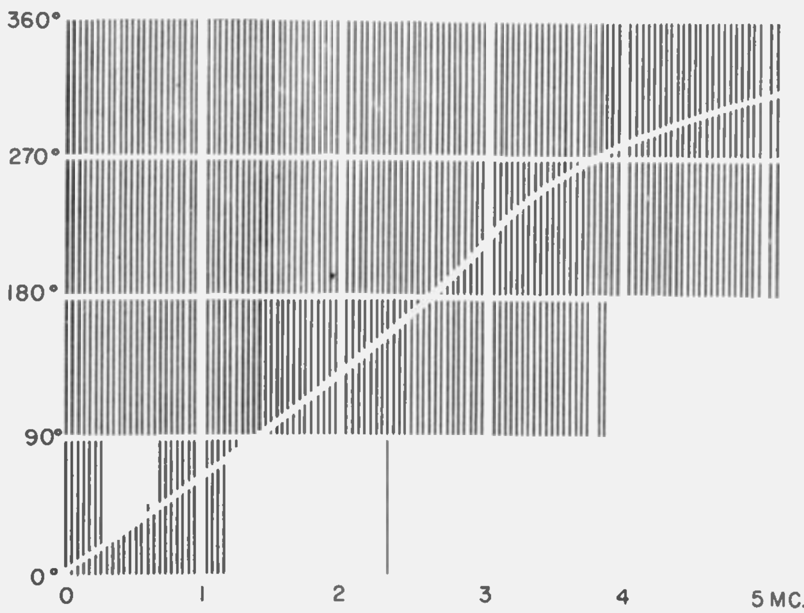


Fig. 9—Traced phase curve.

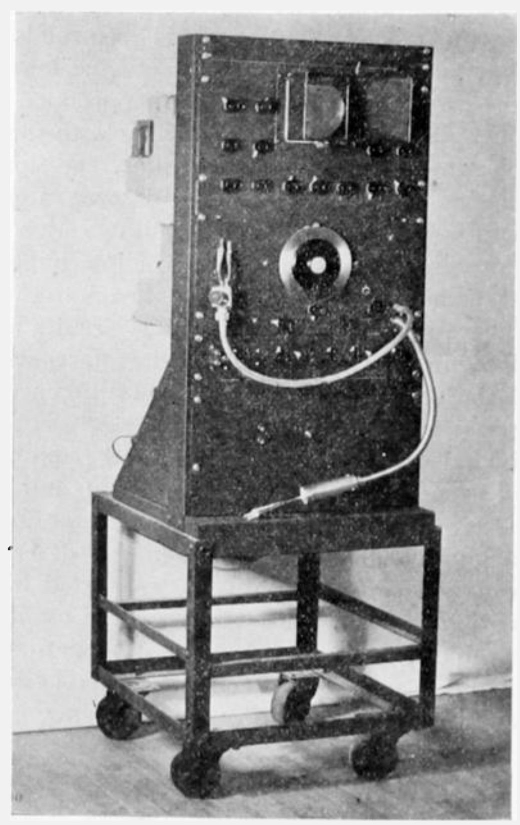


Fig. 10—The phase curve tracer.

plate power to the 67 tubes used in the complete apparatus. In between the middle and bottom units, ananother small panel will be noted. This contains a number of simple circuits that can be used as test circuits for demonstration of the apparatus.

Results

The phase and amplitude characteristics of a number of circuits, as produced on the screen of the cathode-ray tube of this apparatus, are shown in Figs. 11 to 19.

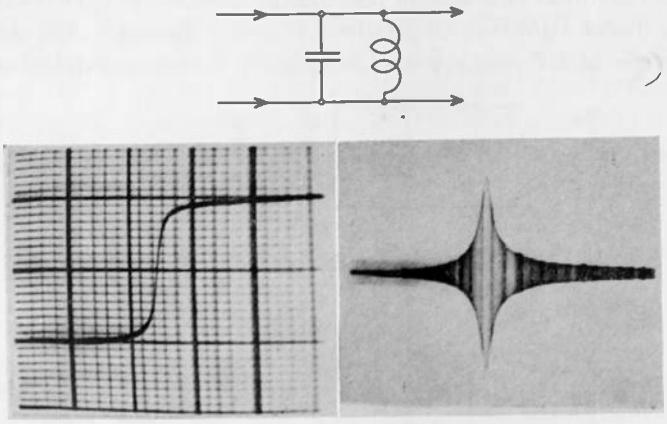


Fig. 11—Single-tuned circuit.

In each case the photograph to the left is the phase characteristic, and to the right the amplitude characteristic. The circuit under test shown in each figure is fed by a constant-current, generator and thus the amplitude and phase characteristics represent the impedance or transfer impedance of the networks.

The characteristics of a single-tuned circuit resonant at about 2.5 megacycles are shown in Fig. 11. The major co-ordinates on the phase curve are at every mega-

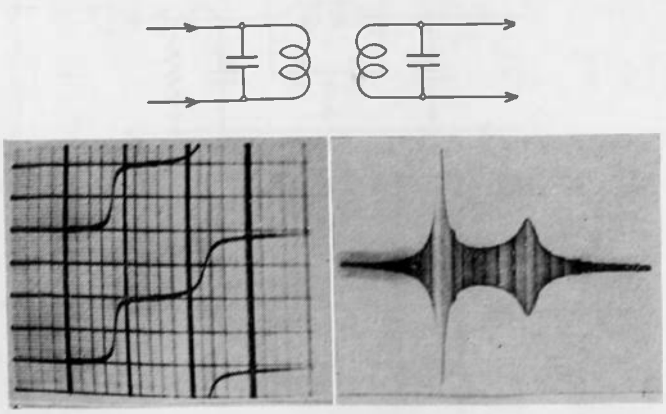


Fig. 12—Double-tuned circuit, overcoupled.

cycle along the horizontal axis and at every 90-degrees along the vertical axis, with minor co-ordinates at every 200 kilocycles and at every 10 degrees. The tuned circuit was in a 1-stage pentode amplifier and, accordingly, at the resonant frequency the phase shift was 180 degrees. The interrelation of a peaked amplitude characteristic and a steep phase slope is indicated; and, as would be expected, a phase shift of ±45 degrees from resonant value occurs at points about 3 decibels down on the amplitude characteristic. The vertical lines on the amplitude characteristic are frequency markers, just as on the phase characteristic.

In Fig. 12 the characteristics of an overcoupled double-tuned circuit are shown. The amplitude characteristic has peaks at about 1.8 and 3.2 megacycles, and the phase curve has "steps" of about 180 degrees

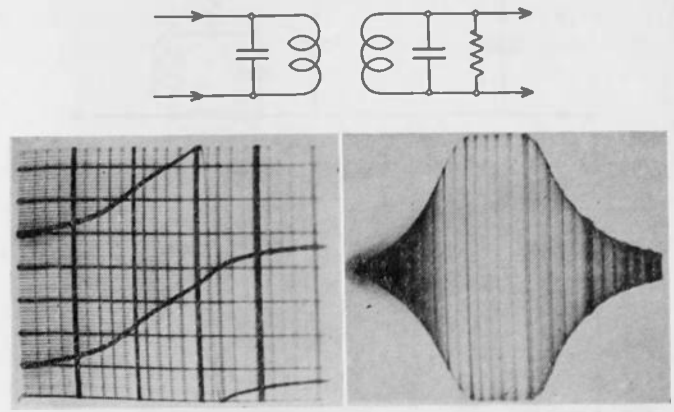


Fig. 13—Double-tuned circuit, optimum-coupled.

(a π step) at the corresponding frequencies. Again, as in Fig. 11, the circuit is fed by a pentode amplifier, so 180 degrees phase shift is added. Thus the phase angle starts at 90 degrees at a low frequency and goes through two π steps as the frequency increases, finally ending up with 450-degree shift at a high frequency. This would require a vertical phase-angle scale of

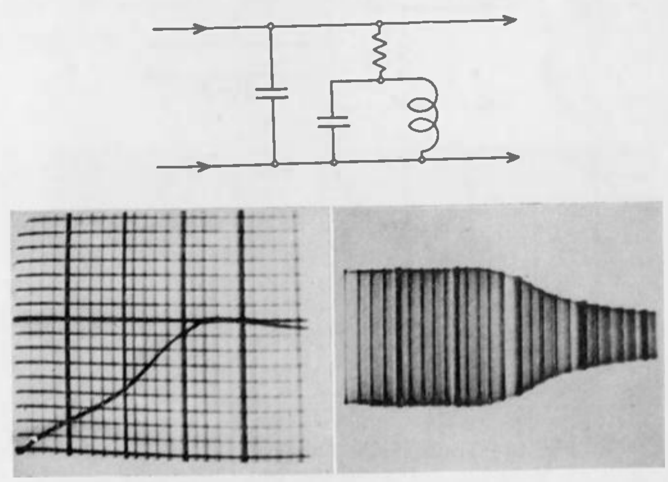


Fig. 14—Half-section filter.

greater than 360 degrees in order to show one continuous phase curve. Accordingly for the phase curve shown in Fig. 12, a 720-degree scale is used. This merely means that the vertical sweep frequency is 25 instead of 50 kilocycles. Since, in general, a phase angle of 360 degrees is not distinguishable from one of 0 degrees, the phase curve is repeated 360 degrees away on the graph. One continuous phase curve is produced, and the other repeated traces may be neglected.

The characteristics of a double-tuned circuit with optimum coupling are shown in Fig. 13. This is the same as Fig. 12, but with the loading increased to that

for which the coupling is optimum, and the signal input to the circuit increased by 20 decibels. The total change in phase angle between 0.1 and 5 megacycles is the same as before, but now the steps in the phase curve are eliminated and the phase slope is almost constant from 1.6 to 3.4 megacycles.

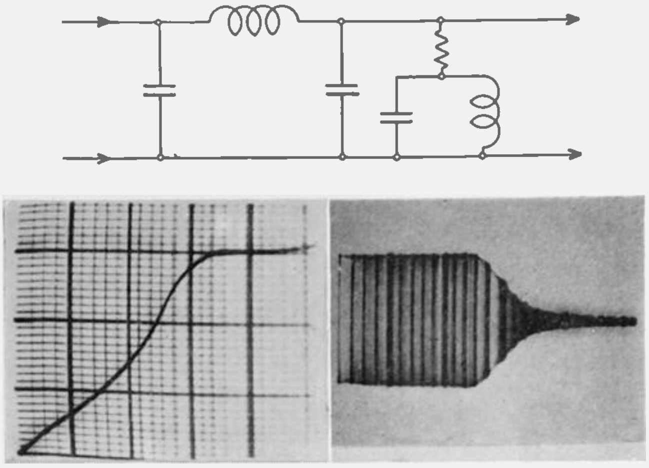


Fig. 15—1½-section filter.

A simple half-section low-pass filter designed for a 3-megacycle cutoff is illustrated in Fig. 14. The vertical scale of the phase curve is expanded by overscanning the cathode-ray tube, so that the major horizontal co-ordinate about halfway up the photograph is 90 degrees. This filter is a two-terminal network that is resistive at a low frequency and predominantly capacitive above cutoff frequency, and thus the total phase shift is 90 degrees. It will be noted that, while the amplitude characteristic is substantially flat out to 2.4 megacycles, the phase curve is not linear over this range, but bends upward indicating that high-frequency components would be delayed more than low-frequency ones.

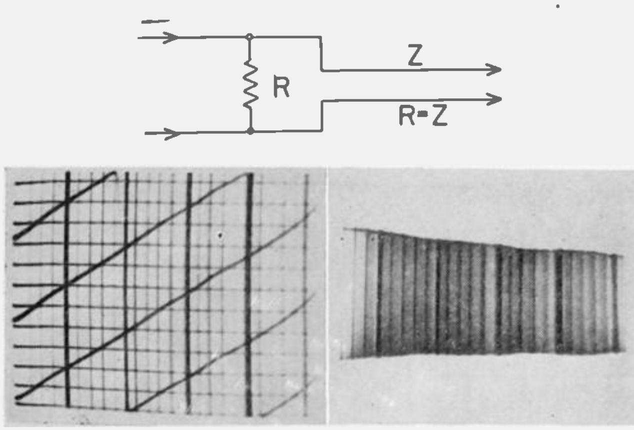


Fig. 16—Transmission line properly terminated at generator end.

Fig. 15 shows the characteristics of a 1½-section filter, and again the cutoff frequency is 3 megacycles. As would be expected, the extra full-section filter adds an extra 180 degrees phase shift up to cutoff frequency, so that 270-degree phase shift exists above cutoff. Again it will be noted that while the amplitude characteristic is substantially flat to 2.6 megacycles, the phase curve is not linear over this range.

The phase and amplitude characteristics of a transmission line are illustrated in Fig. 16. The line consisted of about 250 feet of parallel-wire lamp cord correctly terminated at the generator end and open-circuited at the receiver end. Here the vertical scale on the phase curve is $3\times 360=1080$ degrees, which means that the vertical sweep on the cathode-ray tube is a 16.7-kilocycle linear saw-tooth. The phase shift is substantially a linear function of frequency going through 360 degrees in 2.1 megacycles, indicating a uniform delay of 0.48 microsecond of all frequency components. The amplitude characteristic slopes off slowly as a result of the line attenuation increasing with frequency.

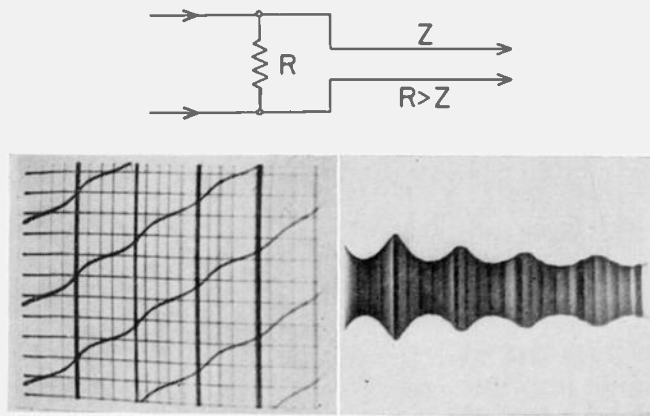


Fig. 17—Transmission line, improperly terminated.

In Fig. 17 we have the same line as in Fig. 16, but the generator resistance is made larger than the surge impedance of the line. The reflections that result from mismatch produce the waves in the amplitude and phase curves. The period of undulation of both the amplitude and phase characteristics is 1.05 megacycles, indicating that these distortions would produce echoes displaced by 0.95 microsecond from the desired output signal. This is the time required for a signal to travel from the receiver end to the generator and back to the receiver again. The maximum slope of the phase curve

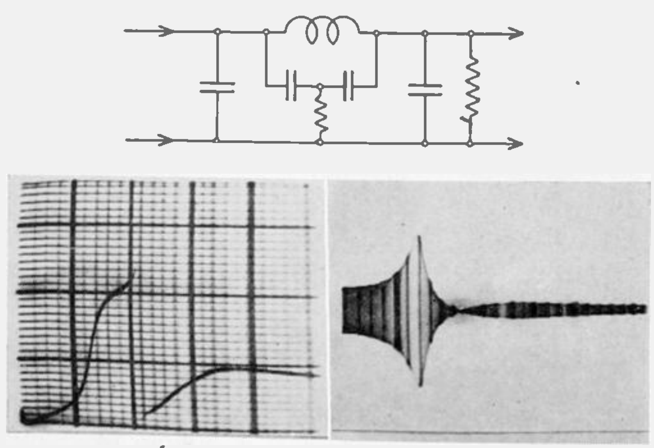


Fig. 18—Trap circuit.

will be seen to occur at the peaks in the amplitude response and the minimum slope at the valleys. It is interesting to note that this circuit has equal amounts of amplitude and phase distortion since, from the nature of the layout, leading echoes are absent and only trailing echoes result.[2] The increasing line attenuation with frequency also attenuates the reflected signals, so the variations in the amplitude and phase characteristics decrease with frequency.

Fig. 18 illustrates the phase and amplitude responses of a high-attenuation trap circuit. At the trap frequency of 2.1 megacycles the phase curve has a downward π step (i.e., a sudden reduction of phase shift by 180 degrees). This is the converse of the upward π step produced when a peak exists in the amplitude curve, as shown by Figs. 11 and 12.

The characteristics of an over-all television system are shown in Fig. 19. The amplitude response is at the top of the figure, and the phase curve is illustrated by the lower photographs, the left photograph having a 360-degree vertical scale and the right a 7×360-degree vertical scale. The system included both a transmitter and receiver. Input was to the monoscope amplifier and output from the last video-frequency stage in a receiver, with the transmitter-receiver connection a radio link on the 50 to 56-megacycle channel. It will be noted that the slope of the phase curve is practically constant up to 2 megacycles, and corresponds to a delay of about 1.2 microseconds. Between 3 and 4 megacycles the phase slope is almost twice the low-frequency value. The phase curve starts at 180 degrees at a low frequency indicating a reversal of picture polarity between the monoscope and picture tube of the receiver. The phase characteristic shown with the 360-degree vertical scale permits accurate measurement of the phase angle at various frequencies, while that shown with the 7×360-degree scale illustrates one continuous phase curve. The amplitude characteristic shows a 4.5-decibel rise at 3.2 megacycles and is down to its low-frequency value at 3.7 megacycles.

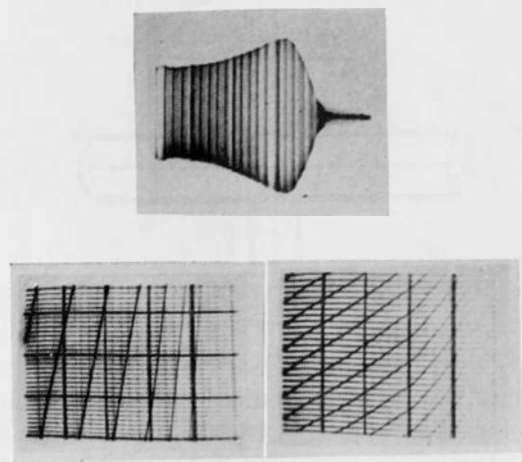


Fig. 19—An over-all television system.

[2] H. A. Wheeler, "The interpretation of amplitude and phase distortion in terms of paired echoes," Proc. I.R.E., vol. 27, pp. 359–385; June, 1939.

A Coaxial Filter for Vestigial-Sideband Transmission in Television*

H. SALINGER,† ASSOCIATE, I.R.E.

Summary—*The problem of building a filter of ladder or lattice type wherein the elements are replaced by coaxial lines is shown to be largely one of geometrical arrangement. A method of designing and constructing such filters is described. Using this procedure, an experimental filter of the ladder type has been built for the television channel of 66 to 72 megacycles. The cutoff sharpness at the lower edge of its frequency range is 32 decibels per per cent of frequency change. This can be achieved with a very compact filter structure. The general performance and range of usefulness of this filter type in television channels is discussed.*

FOR television channels, vestigial-sideband transmission has been generally adopted in this country. Certain standards for this kind of operation have been tentatively proposed and, although they are not yet final, have formed the basis for the work done on this problem in different laboratories. They ask for a very sharp cutoff at the lower end of the transmitter frequency channel. There is also a limitation at the upper end, but this will generally be provided by the video-frequency amplifiers. Therefore, a filter designed to provide the sharp cutoff is essentially a high-pass filter.

Of the possible solutions, the one that will be described here is a filter directly in the carrier-frequency leads between the power amplifier and antenna. The ratio $\Delta f/f'$, where f' is the carrier frequency and Δf the band width within which the transition from the pass band to the attenuation band has to occur, is then of the order of 1 per cent. Obviously, the filter circuits must have as low a dissipation as possible. The best filter elements for this purpose are coaxial transmission lines, and the aim was, therefore, to build the filter entirely out of coaxial sections.

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I. Coaxial Filters

Since filter circuits invariably involve shunt and series connections of different elements, the first question was how to realize these connections in the most efficient manner for coaxial-line elements.[1]

The solution is simple for an element Z shunted across a main line L_1L_2, as shown in Fig. 1. The coaxial equivalent of this is seen in Fig. 2. The element Z,

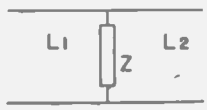

Fig. 1

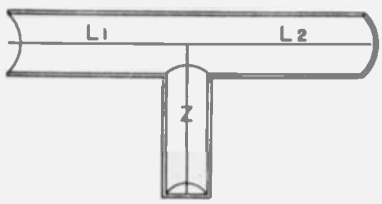

Fig. 2

which may be a short-circuited or an open-end line section, forms a T piece with the main line. At the junction, the input voltage of Z is the same as that of L_2, while the current arriving on L_1 divides itself between L_2 and Z, just as would be the case in Fig. 1.

Let us consider Fig. 3. Z is a line section of which

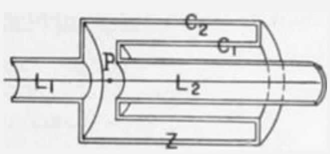

Fig. 3

C_1 forms the inner and C_2 the outer conductor. At point P, the input current of line L_2 is the same as that of Z, while the sum of the input voltages on L_2 and Z equals the voltage at the junction of L_1 and Z. Thus, the equivalent scheme is that of Fig. 4, and Z in this case forms a series element.

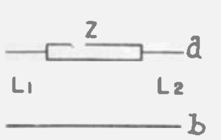

Fig. 4

In Fig. 4, the b conductor contains no impedance elements. Upon comparison, its coaxial equivalent will be found to be the center wire in Fig. 3. This is somewhat unexpected, as it seems more natural to let the outer tube of the main line correspond to the ground conductor of an unbalanced filter mesh. It is possible to turn Fig. 3 inside out so as to meet this condition, but this will generally lead to a more complicated structure with a lower Q, therefore, this possibility will be discarded. It is to be remembered that in Fig. 3 the high-frequency currents are entirely confined to the inside surface of the filter; thus, there is no objection against keeping the outside of L_2 at ground potential as long as direct voltages are excluded.

In Fig. 3, the line Z consists of a radial and an axial portion which meet at right angles. Obviously, the line might also be folded back. An interesting intermediate case is shown in Fig. 5 where the "line" Z takes the

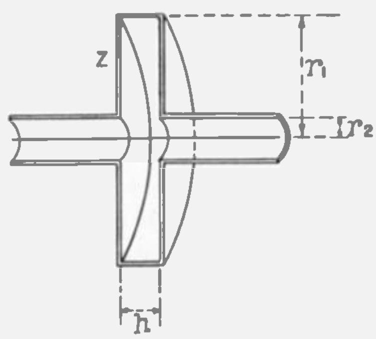

Fig. 5

shape of the space between two circular disks. This structure has been termed a "radial channel." Owing to the fact that its cross section increases continuously as the current travels outward, it shows in itself filter properties, just as an exponential line does. But if it is to be worked near resonance, a very bulky and impractical structure results for frequencies in the present television carrier range. The theory of this arrangement, which is briefly outlined in Appendix I, is, however, useful as it permits the design of the short radial portion of Z in Fig. 3 so as to avoid unwanted reflections at the angle. With the notations in Fig. 5 this radial portion is found to have a characteristic impedance

$$Z_0 = 60h\sqrt{\frac{2 \log r_2/r_1}{r_2^2 - r_1^2}} \tag{1}$$

while its propagation constant $\Gamma = jb$ is given by

$$\sinh \Gamma = j \sin b = j\frac{\omega}{c}\sqrt{\frac{r_2^2 - r_1^2}{2}\log r_2/r_1}. \tag{2}$$

With the shunt and series elements as described, a ladder-type filter may be constructed. It is also possible to include m-derived structures, an example of which will be shown presently.

II. An Experimental Filter

The filter which has really been built is shown in Fig. 6 with Fig. 7 giving the equivalent scheme. Corresponding elements bear the same letter in both figures, and the resonance frequencies of the different parts are also indicated. This arrangement is a two-section filter with m-derivations on both ends. All elements are quarter-wave lines with short-circuited

[1] For previous work along these lines, compare W. P. Mason and R. A. Sykes, "The use of coaxial and balanced transmission lines in filters and wide band transformers for high radio frequencies," *Bell Sys. Tech. Jour.*, vol. 16, pp. 275–302; July, 1937.

ends. In consequence of this, the filter is not of the constant-k m-derived type. A constant-k filter would require some of the elements to be open-ended, which either means radiation losses or, if certain equivalent arrangements are employed, a more complicated structure. Besides, a nonconstant-k filter is just as good as a constant-k one, its only disadvantage being that the mathematical formulas are somewhat more involved.

The short wire which joins G and H in Fig. 7 is in Fig. 6, a piece of coaxial line pqr. This portion is not included in the theory. In order to make pr as short as feasible, this piece of line has a relatively small

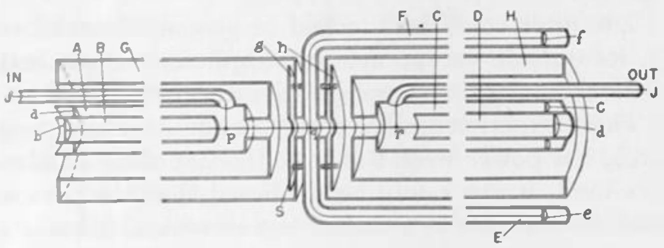

Fig. 6

diameter, with the center wire correspondingly reduced so as to maintain its characteristic impedance. Actually, the distance pr measured 6 inches. All the other connecting lines which appear in Fig. 7 will be found to be negligibly short in the actual arrangement of Fig. 6. This departure from theory made it advisable to make the coaxial elements which constitute the filter tunable. For the elements, A, B, C, D, E, F, this was done by movable plugs a, b, c, d, e, f. Lines G and H are tuned by the two annular disks g and h, which can be moved by screws, thus placing a variable capacitance across the input of G and H. Details of construction are omitted here.

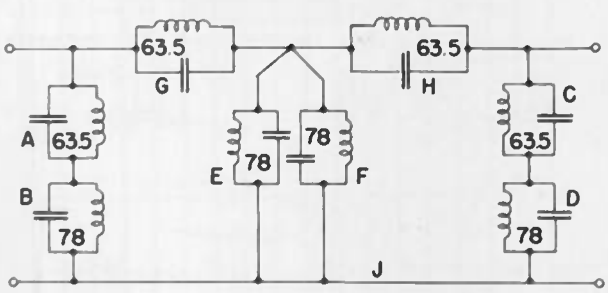

Fig. 7

It is to be understood that Fig. 7 is only approximately equivalent to the real filter. In Fig. 7 the series and shunt impedances would be rational functions of the frequency whereas for the coaxial elements they involve the tangent function. To show how a filter is actually designed, we shall take as an example, instead of the rather complicated structure of Fig. 7, the simple π section of Fig. 8, in which Z_1 and $2Z_2$ are to be embodied by short-circuited transmission lines. We then have

$$Z_1 = jZ_{c1}\,\mathrm{tang}\,\frac{\pi}{2}\frac{f}{f_1}, \quad 2Z_2 = jZ_{c2}\,\mathrm{tang}\,\frac{\pi}{2}\frac{f}{f_2}, \tag{3}$$

where f is the frequency, and the lines are quarter-wave resonant at the frequencies f_1 and f_2, while Z_{c1} and Z_{c2} are their characteristic impedances. Thus, f_1 and f_2 are essentially determined by the lengths of the series and shunt lines; Z_{c1} and Z_{c2}, by the ratios of the tube and center-wire diameters.

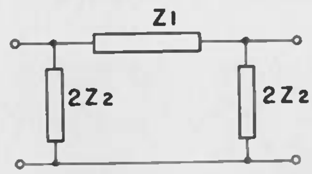

Fig. 8

This π section has itself an iterative impedance Z_0, given by

$$Z_0^2 = \frac{Z_1 Z_2}{1 + Z_1/4Z_2}. \tag{4}$$

It is easily seen that if $Z_1 = 0$ or $Z_2 = 0$ or $Z_2 = \infty$, that is at $f = 2f_1$, $f = 2f_2$, or $f = f_2$, we have a cutoff frequency marking the transition between a pass band and an attenuation band, whereas at $Z_1 = \infty$ ($f = f_1$), Z_0 remains finite and an infinite attenuation peak occurs. Another cutoff frequency, which we shall call f_a, lies at $Z_1 = -4Z_2$. If we use the filter only in the neighborhood of f_1 and f_2, the frequencies where Z_1 or Z_2 vanishes are far outside the working range. We may then choose f_1, f_2, and f_a to lie as in Fig. 9, and the transmission range will extend from f_a to f_2. To determine the four unknown quantities f_1, f_2, Z_{c1}, and Z_{c2}, we may then choose f_1, f_a, f_2, and the value of Z_0 at one frequency between f_a and f_2. Alternatively, if we want essentially only a high-pass filter, f_1, f_a, and the values of Z_0 at two frequencies may be suitably selected in order to determine the four design quantities.

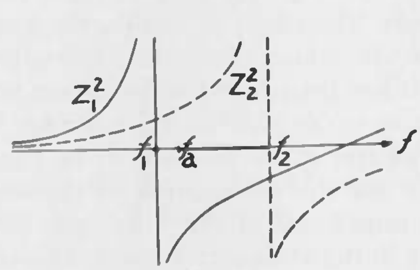

Fig. 9

In either case, if the equation determining f_a

$$Z_1 = -4Z_2$$

is written out explicitly, it will be found that f_a depends not only on f_1 and f_2, but also essentially on the ratio Z_{c1}/Z_{c2}.

The filter was built for the 66- to 72-megacycle television channel. It is constructed from hard brass tubes. Silver-plating the inside surfaces is being contemplated as a possible future development. Fig. 10 shows a view of the filter. It covers a floor space of 13 square feet and takes up a volume of 14.1 cubic feet. The entire

structure, including the steel beam on which it is mounted, weighs about 350 pounds. In weight and bulk, it is believed to be substantially smaller than other filter types used so far. As most television transmitters will probably be located in skyscrapers where space is at a premium, this should be an important advantage of the new construction.

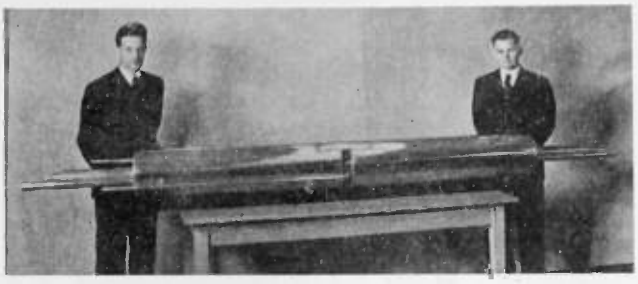

Fig. 10

In Fig. 11, the theoretical curves for the attenuation A of this two-section structure are shown, as well as its characteristic input impedance Z and the delay time T in the pass band. The characteristic impedance on the output end is about 5 per cent higher, as the derivation constants m were chosen slightly different on both ends. Z is seen to be about 123 ohms on the average. In coupling the filter to an antenna, a matching section will have to be interposed. The choice of 123 ohms was made for purely practical reasons. It has been pointed out above that the ratios of characteristic impedances of the filter lines, which means the ratios of tube and wire diameters, are given by the design formulas. Z has then to be chosen so that these diameters remain within practical limits and, moreover, can be realized with standard tube sizes.

The curves in Fig. 11 are computed on a dissipationless basis. The effect of dissipation has been estimated, using Reukema's formulas;[2] but with the actual filter size, it has been found to be almost negligible.

In order to verify the theory, reference is made to the experimental curve marked A_1 in Fig. 11. This, however, is not the attenuation of the filter proper. The input impedance of the filter goes through zero and infinity in the attenuation band. Therefore, it was a simpler matter, and at the same time a closer approximation to actual operating conditions, to excite the grid of the power-amplifier tubes with a constant voltage and to measure the filter output voltage across a load resistor. This curve, of course, includes the amplification in the final stage, and therefore, only the differences in attenuation are relevant, while the total curve has been arbitrarily shifted so as to give $A_1=0$ at its minimum. Also A_1 will include any filtering effect which the tuning and coupling elements in the power-amplifier plate circuits may have. This accounts for the fact that the A_1 values in the attenuation range exceed the theoretical values. In the pass band, there are fluctuations in A_1 by ± 2 decibels, which we ascribe to a mismatch of the load resistor.

Before taking these readings, the filter was tuned up for best performance, but no attention was paid to the actual position of the cutoff, which from Fig. 11 is seen to be too low by about $\frac{1}{3}$ megacycle.

Curve A_1 shows that in the range below 65.2 megacycles the attenuation is at least 39 decibels above its maximum value in the range above 66 megacycles. This means a sharpness of cutoff of 49 decibels per megacycle or 32 decibels per per cent of frequency change.

The filter could be touched or grounded anywhere on its outside except near the input or output ends without having its characteristics affected.

These experiments have been made at a comparatively low power level. But from the size of the conductors used, it can safely be predicted that the present filter will be able to transmit one or several kilowatts.

III. General Remarks

Some explanations are necessary regarding the insertion of such a filter into a television channel. Both its input and output terminals are in the form of transmission lines. Therefore, the output can be connected to the feeder line, if necessary, through a matching section, which will have to be equipped with a quarter-wave insulation piece at its antenna end, in order to prevent high-frequency currents from returning on the outside of the line. Similarly, the filter input end should be extended directly to the transmitter housing so

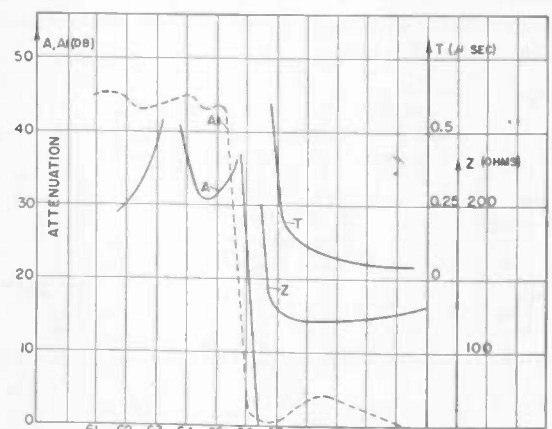

Fig. 11

that no high frequency can creep out and reach the outside of the filter. In its transmission range, the filter presents an ohmic and nearly constant impedance; thus it can be coupled in the conventional manner to a power amplifier. But in the attenuation range, the filter will behave as a reactance and return power into the output tubes. This means that these tubes will have to dissipate the power of the unwanted sideband. However, it can be shown that for amplitude-modulated signals this unwanted power is only a few per cent

[2] L. E. Reukema, "Transmission lines at very high radio frequencies," *Elec. Eng.*, vol. 56, pp. 1002–1011; August, 1937.

of the peak power (see Appendix II). Hence, no serious difficulties are anticipated.

Alternatively, two filters with complementary characteristics may be used, one of which feeds the antenna, while the other one is terminated by a dissipator resistance, a method which already has been used on other filter types.[3] These filters should not be *m*-derived on their input ends, but may be similar in other respects to the one here described.

Another question which arises is whether it is possible to construct a lattice-type filter, using the same principles. Fig. 12 shows one solution of this problem, together with its equivalent structure. The wires at c and d which connect center wires and tubes are, in a certain sense, a departure from the normal use of coaxial lines. But the usual objection that, say at c, a new wave will originate traveling on the outside of the tube Z_2 does not hold in our case, as this wave travels along the line Z_1, and is, therefore, taken account of in the design. The whole structure is completely enclosed.

Finally, it may be asked in which range of frequencies coaxial filters may be useful. Generally speaking, such a filter will decrease in size as the frequency is raised, as it is based substantially on quarter-wave resonant lines. On the other hand, the requirements on the filter become more stringent at higher frequencies, as the band width will become relatively narrower and the sharpness of cutoff expressed in decibels per per cent frequency change will increase with frequency. Thus, it may be necessary to increase the number of filter sections, which is entirely feasible.

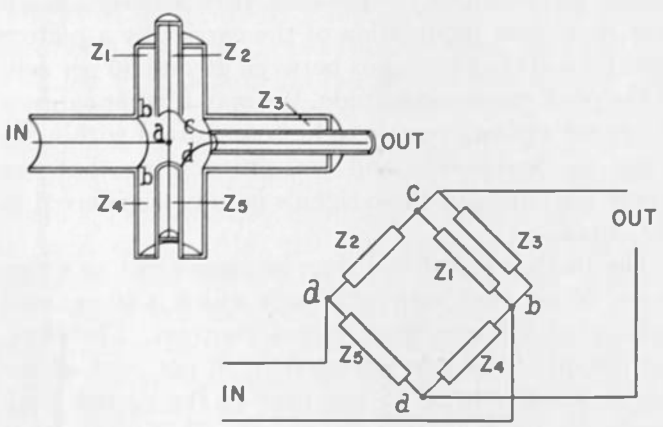

Fig. 12

Referring again to Fig. 7, it is seen that two resonance frequencies, $f_1=63.5$ and $f_2=78$ megacycles, have been used in the present filter for a cutoff at $f_c=66.4$ megacycles. Preliminary calculations have shown that a good design can be obtained throughout the range between 40 and 100 megacycles if the differences f_2-f_c and f_1-f_c are kept at the same values as in this filter.

For frequencies below this range, coaxial filters will have to compete with lumped filter circuits while at higher frequencies, hollow-pipe structures may become an interesting alternative. But within the present television range, it is hoped that the type here described will prove to be a useful tool.

Acknowledgment

Acknowledgment is made to the generous co-operation of the Farnsworth laboratory staff, especially to A. T. Mayle, who helped greatly in solving the constructional problems and in performing the experiments.

Appendix I

Theory of a Radial Channel

In Fig. 13 let i be the current along the disk surface,

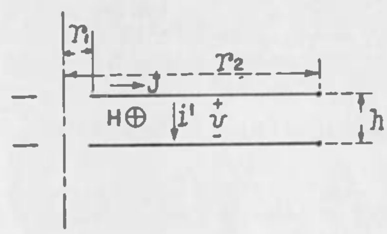

Fig. 13

$i'\cdot dr$ the displacement current between the disks in a ring of width dr; then we have

$$-\frac{\partial i}{\partial r}=i'. \tag{1}$$

With i' there is associated a magnetic field H which is circular around the axis; thus

$$-\frac{\partial}{\partial r}(2\pi rH)=0.4\pi i', \tag{2}$$

where the minus sign is right if H is taken to point away from the reader, as indicated in Fig. 13.

Now let v be the voltage between the disks; Maxwell's induction theorem then gives for sinusoidal currents

$$-\frac{\partial v}{\partial r}=j\omega Hh\cdot 10^{-8}. \tag{3}$$

Finally, the displacement current i' is related with the voltage v by the electrostatic law

$$i'dr=j\omega v\frac{2\pi r dr}{4\pi h}\cdot\frac{1}{9\cdot 10^{11}}, \tag{4}$$

it being, of course, assumed that the disk spacing h is very small compared to the wavelength.

From (2), (3), and (4) the differential equation

$$\frac{\partial^2 v}{\partial r^2}+\frac{1}{r}\frac{\partial v}{\partial r}+\frac{\omega^2}{c^2}v=0 \tag{5}$$

follows, with the general solution

$$v=AJ_0\left(\frac{\omega}{c}r\right)+BY_0\left(\frac{\omega}{c}r\right); \tag{6}$$

J_0 and Y_0 are the Bessel functions of order zero, first

[3] G. H. Brown, "A vestigial side-band filter for use with a television transmitter," *RCA Rev.*, vol. 5, pp. 301–326; January, 1941.

and second kind, respectively. From (1), (2), and (3) we infer

$$i = -j\frac{r}{60h}\left[AJ_1\left(\frac{\omega}{c}r\right) + BY_1\left(\frac{\omega}{c}r\right)\right]. \tag{7}$$

In order to express the integration constants A and B in terms of measurable quantities, let the disks extend from $r=r_1$ to $r=r_2$, and let i_1 and v_1 be the current and voltage at r_1, i_2, and v_2 at r_2. For brevity, we shall write

$$\rho_1 = \frac{\omega}{c}r_1, \qquad \rho_2 = \frac{\omega}{c}r_2. \tag{8}$$

We then readily obtain relations of the familiar type

$$\left.\begin{aligned} v_1 &= \alpha v_2 + \beta i_2 \\ i_1 &= \gamma v_2 + \delta i_2 \end{aligned}\right\} \alpha\delta - \beta\gamma = 1, \tag{9}$$

where

$$\alpha = \frac{\pi}{2}\rho_2[Y_0(\rho_1)J_1(\rho_2) - J_0(\rho_1)Y_1(\rho_2)],$$

$$\beta = 30h\pi j\frac{\omega}{c}[J_0(\rho_1)Y_0(\rho_2) - J_0(\rho_2)Y_0(\rho_1)],$$

$$\gamma = j\frac{\pi}{2}\frac{r_1\rho_2}{60h}[J_1(\rho_1)Y_1(\rho_2) - J_1(\rho_2)Y_1(\rho_1)],$$

$$\delta = \frac{\pi}{2}\rho_1[J_1(\rho_1)Y_0(\rho_2) - Y_1(\rho_1)J_0(\rho_2)]. \tag{10}$$

Equations (9) and (10) contain the entire theory of the radial channel. As is known from network theory, (9) means that we may ascribe two characteristic impedances Z_1 and Z_2 to the channel, depending whether we look into it from the end of r_1 or r_2, and they can be computed from the "nominal impedance"

$$Z_0 = \sqrt{Z_1Z_2} = \sqrt{\beta/\gamma} \tag{11a}$$

and the "transformer ratio"

$$k = \sqrt{Z_1/Z_2} = \sqrt{\alpha/\delta}. \tag{11b}$$

Moreover, there is one propagation constant Γ, which is given by

$$\cosh\Gamma = \sqrt{\alpha\delta} \quad \text{or} \quad \sinh\Gamma = \sqrt{\beta\gamma}. \tag{12}$$

As the expressions (10) vary in sign as ω is varied, it follows that there are pass and attenuation bands for every fixed pair of values r_1, r_2. Conversely, it might be asked how large a channel must be in order to make it antiresonant at a given frequency, if the end is short-circuited. The latter condition means

$$v_1/i_1 = \beta/\delta, \tag{13}$$

thus antiresonance occurs if $\delta=0$ or

$$\frac{J_1(\rho_1)}{Y_1(\rho_1)} = \frac{J_0(\rho_2)}{Y_0(\rho_2)}. \tag{14}$$

Y_0 and Y_1 will be defined as in McLachlan's book.[4] If we take, e.g., $r_1=5$ centimeters, $\omega=2\pi\cdot 60\cdot 10^6$, we find $\rho_1=0.0628$ and the left-hand term of (14) becomes $-3.08\cdot 10^{-3}$. This is so small that we will have to make very nearly $J_0(\rho_2)=0$ or $\rho_2=\omega r_2/c=2.41$. For an ordinary coaxial line we would have $\omega l/c=\pi/2=1.57$ for the same conditions, therefore, r_2 would have to be equal to about $1\frac{1}{2}$ times a quarter wavelength in air.

As pointed out in the paper, a special interest is attached to the case where ρ_1 and ρ_2 are small. In this case (10) reduces to

$$\alpha = 1 + \frac{\rho_2^2-\rho_1^2}{4} - \frac{\rho_2^2}{2}\log\rho_2/\rho_1,$$

$$\beta = 60hj\frac{\omega}{c}\left[\left(1-\frac{\rho_1^2+\rho_2^2}{4}\right)\log\rho_2/\rho_1 + \frac{\rho_2^2-\rho_1^2}{4}\right],$$

$$\gamma = j\frac{\omega}{c}\frac{r_1r_2}{60h}\left[\frac{1}{2}\left(\frac{\rho_2}{\rho_1}-\frac{\rho_1}{\rho_2}\right) + \frac{\rho_1\rho_2}{4}\log\rho_2/\rho_1 + \frac{1}{16}\frac{\rho_1^4-\rho_2^4}{\rho_1\rho_2}\right], \tag{15}$$

while δ is obtained from α by interchanging ρ_1 and ρ_2. In most practical cases, it will be found that the terms in ρ^2 can also be neglected; thus $\alpha=\delta=k=1$ and $Z_0=Z_1=Z_2$. This leads immediately to (1) and (2) of the paper.

Appendix II

Power Contained in the Suppressed Sideband

Let us take the usual case of negative-polarity transmission. The peak carrier amplitude for synchronizing signals will be called 100 per cent, then we may assume that maximum modulation of the carrier by a picture signal results in excursions between 20 and 80 per cent of the peak carrier amplitude. We may further suppose the synchronizing signals to be transmitted within the range of double-sideband transmission so that the power contained in these signals is fully transferred to the antenna.

The picture signal will then be equivalent to a carrier of 50 per cent peak amplitude which is 60 per cent modulated for maximum signal contrast. Therefore, the amplitude of one sideband is 30 per cent of the carrier amplitude or 15 per cent of the carrier peak amplitude for synchronizing signals. This means that the power per sideband is only 2.25 per cent of the peak power developed while a synchronizing signal occurs, and 3.5 per cent of the power for continuous operation at black level (equals 80 per cent of peak amplitude).

Various other assumptions may be made, but it will always be found that the unwanted power is only a few per cent of the maximum power to be delivered by the transmitter, as long as amplitude modulation is considered.

For frequency modulation, this will no longer be true, and we get large sideband power. In this case, the filter design has to be modified as pointed out in section III of this paper.

[4] N. W. McLachlan, "Bessel Functions for Engineers," Oxford University Press, Oxford, England, 1934.

On the Theory of Tubes with Two Control Grids*

ALEXANDER H. WING†, ASSOCIATE, I.R.E.

Summary—*The plate current at constant screen and plate voltages in a tube having two well-shielded control grids acting in tandem on the same electron stream is shown to be equal to the product of two functions, each a function of the potential on one grid only. When the second control grid has a uniform structure, the plate current is practically zero for second-control-grid voltages more negative than a value obtained by dividing the screen voltage by a suitable μ factor.*

For the pentagrid mixer hexode as modulator and heterodyne detector, simplified equivalent circuits are shown to hold, whereby the desired-frequency output currents and voltages may be obtained by calculations similar to those for class A amplifiers. These calculations may be quickly and accurately made. The timesaving is great since a multiplicity of series terms need not be calculated.

Where the outer control grid has a uniform structure, the outer pentode has a definite cutoff amplification factor.

In terms of the grid-voltage functions and measurements of plate current and plate resistance at known control-grid voltages, the performance of the tube may be calculated when the grid voltages are known.

For the pentagrid mixer hexode as modulator and detector, simplified equivalent circuits similar to those of an amplifier are shown to hold. A method is given whereby the desired-frequency output currents may be obtained by calculations involving simple trigonometric functions. These calculations may be quickly made and yield accurate results. The timesaving is great since a multiplicity of series terms is not involved.

INTRODUCTION

THERE ARE many multielement tubes in which there are two control grids acting in tandem on a single electron stream. Though they are capable of very wide application, they are commonly used as amplifiers whose gain can be controlled for purposes of volume contraction, volume expansion, and volume limiting; as modulators; and as heterodyne detectors. The development of these tubes came after the development of the pentode. Each tube was usually developed with a specific application in mind, and the appearance of each tube on the market was followed by the appearance in the literature of articles describing the tube in the particular application for which it was intended. However, many of these articles have the common fault of being qualitative rather than quantitative in nature. Also, the data published by the tube manufacturers have in general been rather meager. Such published data are usually confined to a few charts and a set of "typical operating conditions," so that if the tube is operated at other than a specified set of voltages, the calculation of the tube performance is virtually impossible. It is the purpose of this paper to describe certain characteristics of these tubes, which previously have not been described, and to set forth a method based upon these characteristics whereby the performance of these tubes can be quantitatively evaluated by methods which are not so time-consuming as to prevent their use in practice.

The specific data here presented concern the pentagrid heptode,[1,2] whose type designation is 6L7 for the metal envelope and 6L7G for the glass envelope. This tube was selected because it was designed as a single-purpose tube in which each control grid has the single function of controlling the plate current only, and the two control grids are well shielded from each other and the plate. Also the tube is cylindrical in structure so that except for leakage effects at the edges of the cylindrical grids, the entire space current is under the control of both control grids. The type 6L7 is commonly called a "pentagrid mixer" since five of its elements are grids, and it was developed to be used mainly as a heterodyne detector or "mixer" in which signals of two different frequencies are combined or mixed to produce a signal having a third frequency equal to the difference between the first two frequencies.

ANALYSIS OF TUBE CHARACTERISTICS

The arrangement of the electrodes of the 6L7 is indicated in Fig. 1. The cathode is circular in shape. The grids, numbered in order of their distance from the cathode, are termed grids 1, 2, 3, etc. Grid 1 is oval in shape with nonuniform spacing of grid wires.[3] Because of the oval shape, the radial distance between the cathode surface and grid 1 is not the same in all directions. Grid 2 is also oval in shape but has uniform spacing of grid wires. The oval shapes of grids 1 and 2 are such that the radial distance between grids 1 and 2 is not the same in all directions. Grids, 3, 4, and 5 are circular in shape with uniform spacing of grid wires. The plate is circular. Grids 1 and 3 are the control grids. Grids 2 and 4 are shield or screen grids. Grid 5 is a suppressor grid connected internally to the cath-

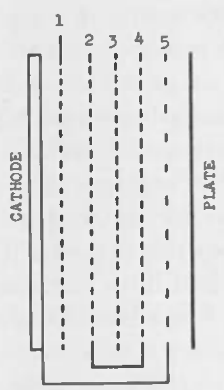

Fig. 1—Arrangement of electrodes type 6L7 tube.

* Decimal classification: R132×R139. Original manuscript received by the Institute, November 4, 1940. This paper is a dissertation submitted in partial fulfillment of the requirements for the degree of doctor of philosophy in the faculty of pure science of Columbia University in the City of New York.

† College of the City of New York, School of Technology, New York, N. Y.

[1] C. F. Nesslage, E. W. Herold, and W. H. Harris, "A new tube for use in superheterodyne frequency conversion systems," PROC. I.R.E., vol. 24, pp. 207–218; February, 1936.

[2] Application Note No. 50, The operation of the 6L7 as a mixer tube, RCA Manufacturing Co., Inc., August 19, 1935.

[3] In Fig. 1 of reference (2) a sketch is given in which grids 1 and 2 are represented as being circular. The actual structural shape is as here stated.

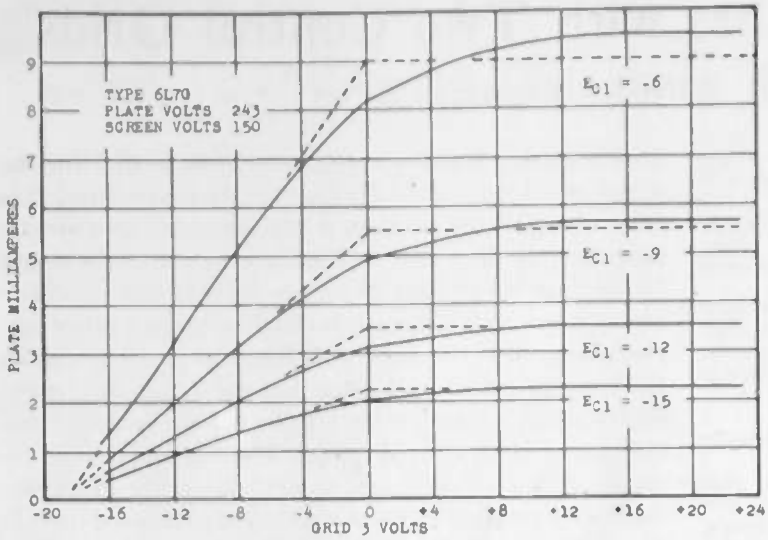

Fig. 2—Effect of grid-3 voltage on plate current at various grid-1 voltages.

ode. The two screen grids, 2 and 4, are connected together internally to the same base pin. This is done to reduce the number of connections brought out from the tube and is not essential for the operation of the tube. Because of this connection, however, screen grids 2 and 4 are constrained to operate at the same potential. The plate and grid 3 are carbonized to prevent secondary emission.

In Fig. 2 there is presented a family of curves of plate current (i_p) as ordinate, with grid-3 voltage (e_{g3}) as abscissa, and with grid-1 voltage (e_{g1}) held constant at the various bias voltages (E_{C1}) indicated. These curves have almost identical shape. Table I presents the data from which the curves of Fig. 2 were plotted. In Table II are found data calculated column by column from Table I. The value of i_p given in each column of Table I for $E_{C3}=+6$ was taken as unity and the other values of i_p in the same column were calculated in terms of this unit. It is seen from the data in Table II that at constant voltage on grid 1 the percentage variation in i_p as e_{g3} is varied is practically independent of the voltage on grid 1, since the difference between similarly located numbers in the columns is less than 0.05 in almost all cases. This indicates that the form of the function expressing the plate current in terms of grid-3 voltage is independent of the voltage on grid 1.

Since a change in grid-1 voltage alters the size but not the shape of the curves in Fig. 2, these curves may be expressed by the relation[4]

$$i_p = if_3(e_{g3}) \tag{1}$$

where i is the plate current at $e_{g3}=+6$ and $f_3(e_{g3})$ is a function of grid-3 voltage determined by plotting the numbers in Table II as a function of e_{g3}. This plot is given in Fig. 3.

Now, i in (1) is a function of the voltage on grid 1. The form of this function may be obtained in a manner similar to that used in determining $f_3(e_{g3})$. In Table III are found data computed row by row from Table I, by taking the value of i_p at $E_{C1}=-3$ as unity and expressing the other values of i_p in the same row in terms of this unit. Here again, the tabular values indicate a remarkably similar variation. Except for grid-3 voltages more negative than 12 volts, the difference between similarly located numbers in the table is less than 0.03. This indicates that the functional form interrelating the plate current with the grid voltage e_{g1} is, over very wide limits, independent of the voltage on grid 3. This functional form may be determined by plotting the numbers of Table III as a function of e_{g1}. This is done in Fig. 4. Therefore, if I represents the value of i in (1) at $e_{g1}=-3$, then

$$i = If_1(e_{g1}) \tag{2}$$

where $f_1(e_{g1})$ is the function of Fig. 4.

Substitution of (2) into (1) results in

$$i_p = If_1(e_{g1})f_3(e_{g3}). \tag{3}$$

This relationship has not been pointed out in previous discussions of tubes with two control grids, and forms the basis of this paper.

Equation (3) states that the plate current of a tube having two control grids acting in tandem on the same electron stream, in which the outer control grid is

TYPE 6L7G $E_P=243$ $E_{B1}=150$

TABLE I

I_P MILLIAMPERES

E_{C3}	E_{C1} −3	E_{C1} −6	E_{C1} −9	E_{C1} −12	E_{C1} −15
+22.5	15.8	9.50	5.66	3.53	2.26
+18	15.8	9.50	5.70	3.55	2.26
+12	15.6	9.43	5.68	3.55	2.27
+ 6	14.9	9.05	5.44	3.42	2.22
+ 3	14.2	8.67	5.18	3.27	2.12
0	13.5	8.19	4.92	3.13	2.02
− 2	12.4	7.59	4.54	2.88	1.90
− 4	11.2	6.87	4.11	2.64	1.75
− 6	9.9	6.06	3.67	2.37	1.57
− 8	8.5	5.19	3.15	2.04	1.36
−10	6.85	4.20	2.60	1.68	1.14
−12	5.2	3.27	2.00	1.32	0.90
−14	3.42	2.24	1.41	0.93	0.63
−16	2.02	1.39	0.88	0.60	0.41
−18	1.05	0.72	0.50	0.35	0.25
−20	0.41	0.31		0.16	0.12

TABLE III

I_P AS A FUNCTION OF E_{C1}
I_P AT $E_{C1}=-3=$ UNITY

E_{C1} −3	E_{C1} −6	E_{C1} −9	E_{C1} −12	E_{C1} −15
1.00	0.60	0.36	0.22	0.14
1.00	0.60	0.36	0.23	0.14
1.00	0.60	0.36	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.36	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.38	0.26	0.17
1.00	0.63	0.38	0.25	0.17
1.00	0.66	0.41	0.27	0.18
1.00	0.69	0.44	0.30	0.20
1.00	0.69	0.48	0.33	0.24
1.00	0.76		0.39	0.29

TABLE II

I_P AS A FUNCTION OF E_{C3}
I_P AT $=E_{C3}=+6=$ UNITY

+22.5	1.06	1.05	1.04	1.03	1.02
+18	1.06	1.05	1.05	1.04	1.02
+12	1.05	1.04	1.05	1.04	1.02
+ 6	1.00	1.00	1.00	1.00	1.00
+ 3	0.95	0.96	0.95	0.96	0.96
0	0.91	0.90	0.90	0.92	0.91
− 2	0.83	0.84	0.83	0.84	0.86
− 4	0.75	0.76	0.76	0.77	0.79
− 6	0.66	0.67	0.67	0.69	0.71
− 8	0.57	0.57	0.58	0.60	0.61
−10	0.46	0.46	0.48	0.49	0.51
−12	0.35	0.36	0.37	0.39	0.41
−14	0.23	0.25	0.26	0.27	0.28
−16	0.14	0.15	0.16	0.18	0.18
−18	0.07	0.08	0.09	0.10	0.11
−20	0.03	0.03		0.05	0.05

[4] The list of symbols will be found in Appendix II.

shielded on both sides, ıay be expressed as a constant times the prodct of two functions, each function involving te voltage on one control grid only. The constnt is the value of the plate current at the gri voltages at which both of the functions ha unity value.

Since the 6L7 and othr tubes having more than one control grid hae these grids as well as the screen grids intrposed between the plate and the cathode, ıe plate is very well shielded from the cathod In this respect the 6L7 is similar to a well-shlded pentode. In the 6L7, as in a well-shieldd pentode, the plate voltage has but a minor nfluence on the plate current. Its chief effect iso change slightly the value of I in (3). To tak account of this, we

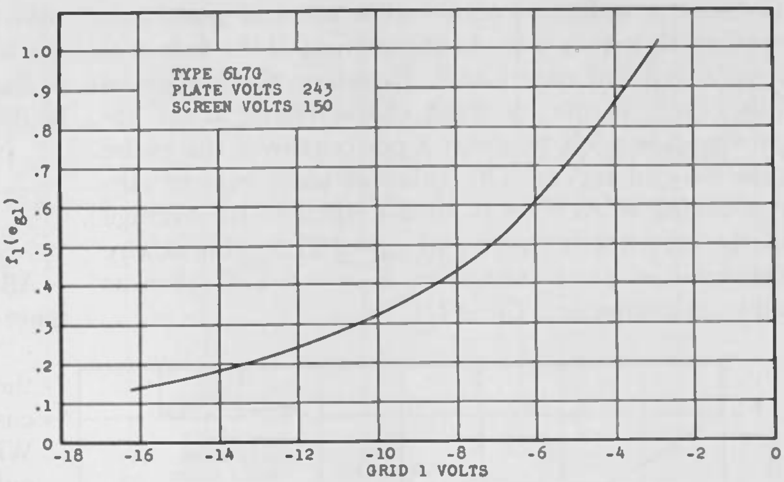

Fig. 4—Plate current as a function of grid-1 voltage expressed in terms of value at $E_{C1} = -3$ volts.

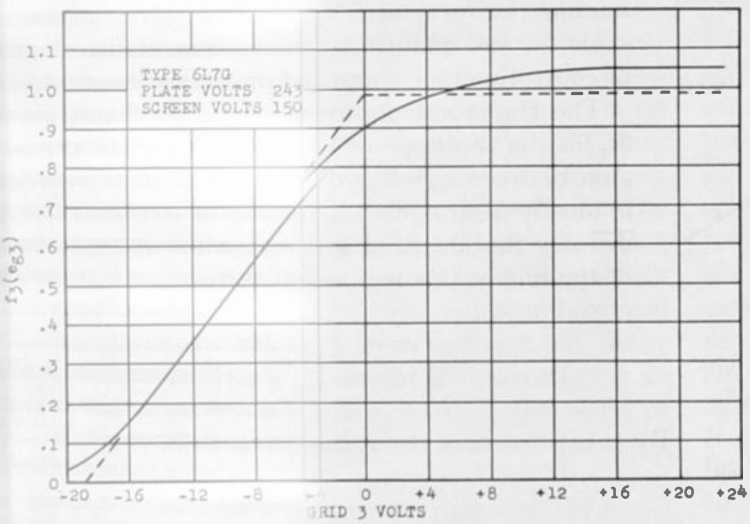

Fig. 3—Plate curnt as a function of grid-3 voltage expressed in tens of value at $E_{C3} = +6$ volts.

may write

$$I = I_0 f_p(e_p) \tag{4}$$

where I_0 is the value of I at the normal plate-bias voltage. For all plate ltages considerably above the screen voltag the plate current is practically independent the plate voltage, so that $f_p(e_p)$ has a value ry close to unity in this range.

The screen voltage is aındependent and arbitrarily fixed parametervolved in all of the tube characteristics. Its eect on $f_3(e_{g3})$, $f_1(e_{g1})$, and $f_p(e_p)$ will now be shrn.

In Fig. 5 plate currents plotted as a function of e_{g3} at a fixed voltae on grid 1, for various values of screen volge. Here again the general form of the curv is the same as that in Figs. 2 and 3. This forı is such that it may to a very close approximtion be represented by three straight lines, otained in the following manner. First, a diaonal line is drawn practically to coincide wh the most linear part of the curve in the egative region, with little weight given to the roundness of the curve in the region near zero grid voltage. Second, from the intersection of this diagonal line with the axis of ordinates, a horizontal straight line is drawn extending into the region of positive grid voltage. From the intersection of the diagonal line with the axis of the abscissas, a third line is drawn along the axis of the abscissas, indicating a region in which the plate current is zero. These idealized characteristics have been indicated in Figs. 2, 3, and 5 by the dashed lines.

The idealized characteristic straight line in the positive-grid region is suggested by the flatness of the curves of Figs. 2, 3, and 5 in this region and by the fact that there is an actual saturation current which is not exceeded, no matter how positive grid 3 is driven. The plate current at grid-3 voltages causing the plate current to reach saturation (a value $e_{g3} = +22.5$ is more than sufficient) and the plate current at $e_{g3} = 0$ are not very different, the difference being about 1.2 decibels or 15 per cent. The intersection of the diagonal line with

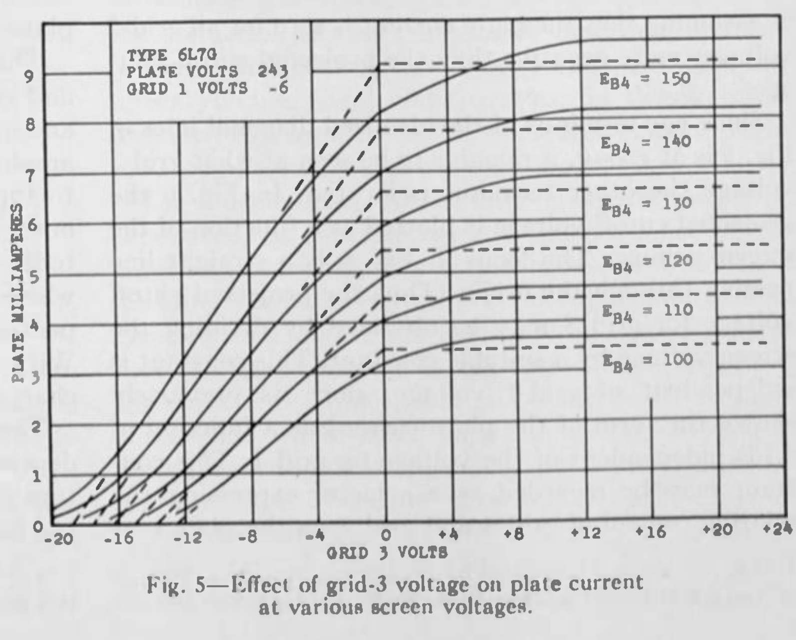

Fig. 5—Effect of grid-3 voltage on plate current at various screen voltages.

the axis of ordinates occurs at a value of plate current which is very close to the average of the values of i_p at $e_{g3}=0$ and $e_{g3}=+22.5$. Therefore, the horizontal line taken as the idealized characteristic is an approximation good to about 8 per cent over the entire positive-grid region. The value of plate current corresponding to $e_{g3}=+6$ is almost equal to the average of the currents at $e_{g3}=0$ and $e_{g3}=+22.5$. This is why the value of plate current at $E_{C3}=+6$ was taken as unity in computing Table III.

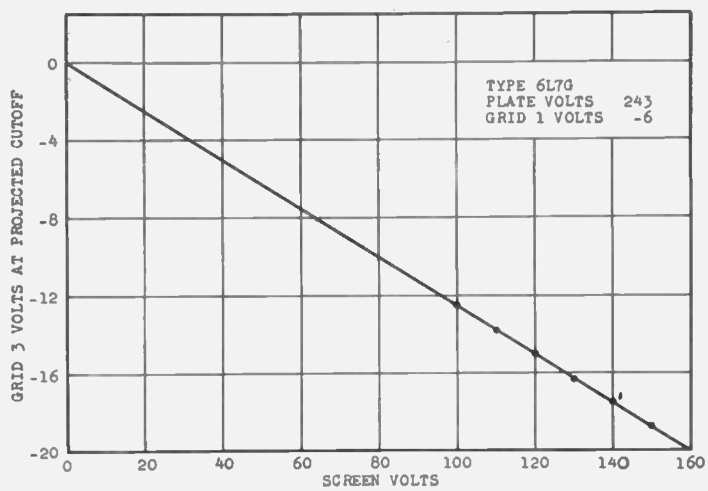

Fig. 6—Effect of screen voltage on grid-3 cutoff voltage.

The intersection of the diagonal line with the axis of the abscissas establishes a voltage which will be termed the projected cutoff voltage to distinguish it from the actual voltage required to produce actual plate-current cutoff. The actual cutoff voltage cannot be sharply defined because of the effects of irregular tube structure,[5] electron leakage around the ends of the grid structures, contact differences of potential, and initial velocity of emission of electrons from the cathode. The very small plate currents in the region near cutoff contribute practically nothing to the performance of the tube, so that there is negligible error in assuming that the plate current is zero for all grid-3 voltages more negative than the projected cutoff voltage.

Since one terminus of the idealized diagonal lines of Fig. 5 is at $e_{g3}=0$, it remains to be seen at what grid-3 voltage the other terminus is located. In Fig. 6 the projected cutoff voltage is plotted as a function of the screen voltage. The locus in Fig. 6 is a straight line passing through the origin. Thus the projected cutoff voltage for grid 3 may be obtained by dividing the screen voltage by a suitable constant. This constant is independent of grid-1 voltage, since as previously shown the form of the plate current as a function of e_{g3} is independent of the voltage on grid 1. This constant may be regarded as a μ factor expressing the relative control of grid 4 and grid 3 on the plate current. Thus, if the screen is biased to a voltage E_{B4}, and the bias required on grid 3 to produce projected cutoff is E_{3CO}, then the μ factor between grids 3 and 4 may be defined by the equation

$$\mu_{34} = -\frac{E_{B4}}{E_{3CO}}. \tag{5}$$

All voltages are taken in their algebraic or vector sense or direction with respect to the cathode.

The relationship of (5) is not even hinted at in the published data on the 6L7 tube, nor is it discussed in articles describing uses of the 6L7 tube. Without knowledge of this relationship, one can only guess as to the proper bias and control voltages for grid 3 in the many circuits to which the tube is adaptable. Equation (5) is vital in determining the form of $f_3(e_{g3})$ and therefore indispensable for the quantitative evaluation of the effects of control voltages impressed on grid 3.

The transconductance between grid 3 and plate $\partial i_p/\partial e_{g3}$ is the slope of the i_p-e_{g3} curve. In the region between $e_{g3}=0$ and $e_{g3}=E_{3CO}$ the plate current is ideally a straight-line function of e_{g3}; then if i_{PS} denotes the idealized saturation current, the slope of the line in this region is given by

$$g_{3P}\Big]_{e_{g3}=E_{3CO}}^{e_{g3}=0} = \frac{i_{PS}}{|E_{3CO}|}. \tag{6}$$

By substitution of the value of E_{3CO} from (5),

$$g_{3P}\Big]_{e_{g3}=E_{3CO}}^{e_{g3}=0} = \frac{i_{PS}\mu_{34}}{E_{B4}}. \tag{7}$$

In the region where e_{g3} is more negative than E_{3CO}, the transconductance between grid 3 and the plate is ideally zero because of plate-current cutoff, and in the region where e_{g3} is positive, the transconductance between grid 3 and the plate is ideally zero because of plate-current saturation.

The relationship between plate current and screen and grid-3 voltages as disclosed by Figs. 2, 3, 5, and 6 and by (5) is exactly the same as in an ordinary voltage-amplifier pentode. The μ factor of (5) is analogous to the cutoff amplification factor of a pentode.[6] The outer portion of the heptode is, then, virtually a pentode. There is virtually a cathode in front of grid 3 whose emission limits the plate current of the virtual pentode to the saturation currents of Figs. 2 and 5. With respect to grid 3, grids 1 and 2 act as space-charge grids fixing the emission of the virtual cathode.

The emission of the virtual cathode just described depends upon the voltages on grids 1 and 2. The nature of the control exerted by grid 1 (a control grid) at a fixed voltage on grid 2 (a screen grid) is indicated in

[5] E. L. Chaffee, "Theory of Thermionic Vacuum Tubes," McGraw-Hill Book Co., New York, N. Y., 1933, pp. 180–190.

[6] F. E. Terman, "Radio Engineering", second edition, McGraw-Hill Book Co., New York, N. Y., 1937, p. 140.

Fig. 4. This control is not linear in this tube (6L7) since grid 1 is a variable-μ grid. This variable-μ characteristic is caused primarily by the variable pitch of the grid wires, and secondarily by the fact that grids 1 and 2 are oval in shape so that the radial distances of the grid wires from the cathode are not the same in all radial directions. Further, the support rods for grid 1 have a diameter which is an appreciable fraction of even the greatest distance between grid 1 and the cathode. Thus the electrode configuration of grids 1 and 2 is that of a triode, but because of structural nonuniformity in both radial and axial dimensions, the effects are similar to what would be obtained if many smaller triodes were connected in parallel, each triode having a different μ factor[5] for grid 1. As grid 1 is made more negative, the high-μ triodes cut off their plate current first, leaving the low-μ triodes operative. Further, if the screen voltage were lowered, the high-μ triodes would cut off sooner. Therefore, there is no simple function whereby the effects of grid 1 may be described, because when either grid-1 voltage or screen voltage is changed, the effect is as though the structure of the tube were changed. However, it can be definitely stated that at lower (positive) screen voltages, the high-μ parts of the grid structure will cut off at lower (negative) grid voltages. At lower screen voltages, the plate current decreases faster as grid 1 is made more negative, so that the slope of $f_1(e_{g1})$ at a given value of e_{g1} is increased when the screen voltage is decreased.

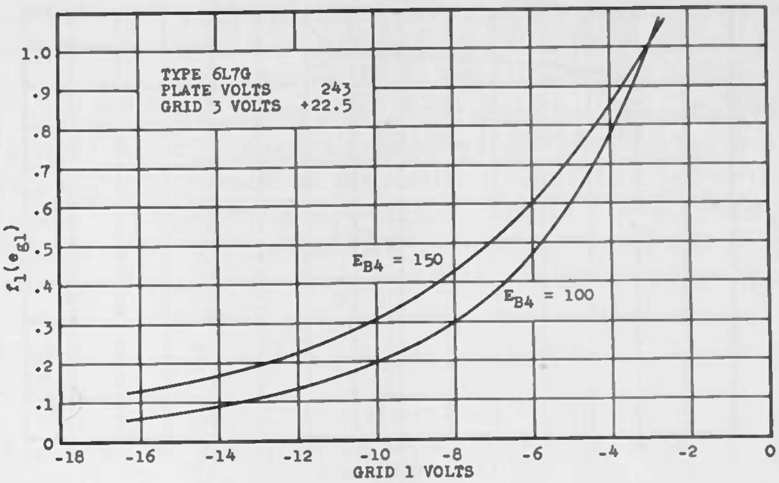

Fig. 7—Plate current as a function of grid-1 voltage at various screen voltages.

In Tables IV and V data and computations of $f_1(e_{g1})$ for screen voltages between 150 and 100 volts are given. Plots of $f_1(e_{g1})$ for screen voltages of 150 and 100 are given in Fig. 7. These curves bear out the conclusions of the previous paragraph.

TYPE 6L7G $E_P=243$ $E_{C3}=+22.4$

TABLE IV

I_P MILLIAMPERES

E_{P4}	E_{C1} −3	E_{C1} −6	E_{C1} −9	E_{C1} −12	E_{C1} −16
+150	15.6	9.4	5.6	3.5	1.95
+140	13.8	8.1	4.7	2.9	1.51
+130	12.3	6.85	3.87	2.31	1.19
+120	10.4	5.46	2.99	1.72	0.88
+110	8.91	4.34	2.37	1.13	0.65
+100	7.50	3.53	1.80	0.95	0.48

TABLE V

I_P AS A FUNCTION OF E_{C1}

I_P AT $E_{C1}=-3=$ UNITY

+150	1.00	0.60	0.36	0.22	0.13
+140	1.00	0.59	0.34	0.21	0.11
+130	1.00	0.56	0.31	0.19	0.10
+120	1.00	0.53	0.29	0.17	0.08
+110	1.00	0.49	0.27	0.13	0.07
+100	1.00	0.47	0.24	0.13	0.06

It has been shown above that the behavior of a tube with two control grids in tandem, having a screen grid in between the control grids and a screen grid and a suppressor grid in between the second control grid and plate, is like that of a triode in tandem with a pentode. The first grid and screen form an inner triode on whose action grid 1 has a controlling influence as in an ordinary triode. The electrons of the space current of this triode not taken up by the first screen pass through to form a virtual cathode in front of the second control grid; this virtual cathode in conjunction with the second control grid, the second screen grid, the suppressor, and the plate, form a pentode on whose action the second control grid has a controlling influence as in an ordinary pentode having an actual cathode. The plate current as controlled by the two control grids may be expressed as the product of two independent functions times a constant, the constant being the plate current corresponding to those control-grid voltages for which the respective functions have unity value. The function describing the action of the first control grid depends upon the first screen voltage and the function describing the action of the second control grid depends on the second screen voltage. When the control grid has a variable-μ structure, the control action of this grid cannot be described in terms of a μ factor. When the control grid has a constant-μ structure, then the control action of this grid can be described in terms involving this constant μ factor. In any event, at any screen voltage, $f_1(e_{g1})$ and $f_3(e_{g3})$ can be definitely determined and from these functions the behavior of the tube may be quantitatively evaluated.

The independence of the functions $f_1(e_{g1})$ and $f_3(e_{g3})$ depends upon the electrostatic independence of the inner and outer sections of the tube. With sufficient shielding, the outer control-grid potential cannot affect the potential distribution within the screen grid so that the space current in the region between grid 2 and cathode is a function of grid-2 and grid-1 voltages only. Under these conditions the virtual emission of the virtual cathode in front of grid 3 is independent of grid-3 voltage. Therefore, (3) will apply where this independence obtains.

The form of $f_p(e_p)$ is indicated in Fig. 8, where curves of $f_p(e_p)$ for screen voltages of 100 and 150 volts

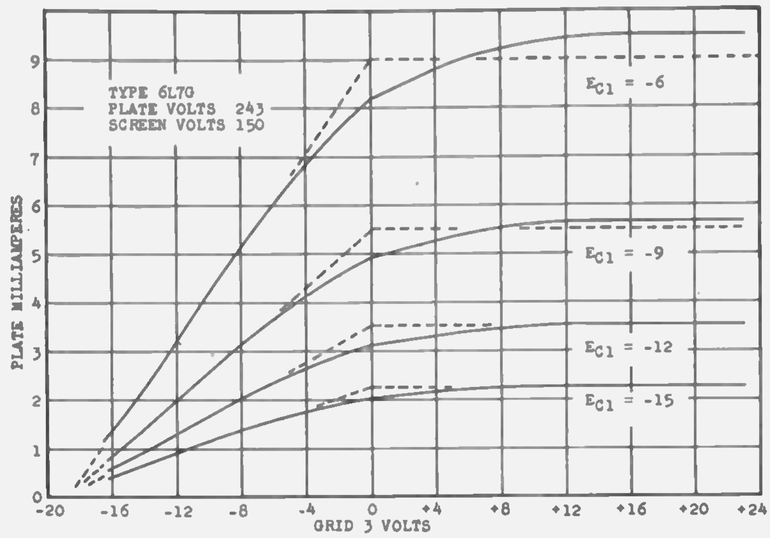

Fig. 2—Effect of grid-3 voltage on plate current at various grid-1 voltages.

ode. The two screen grids, 2 and 4, are connected together internally to the same base pin. This is done to reduce the number of connections brought out from the tube and is not essential for the operation of the tube. Because of this connection, however, screen grids 2 and 4 are constrained to operate at the same potential. The plate and grid 3 are carbonized to prevent secondary emission.

In Fig. 2 there is presented a family of curves of plate current (i_p) as ordinate, with grid-3 voltage (e_{g3}) as abscissa, and with grid-1 voltage (e_{g1}) held constant at the various bias voltages (E_{C1}) indicated. These curves have almost identical shape. Table I presents the data from which the curves of Fig. 2 were plotted. In Table II are found data calculated column by column from Table I. The value of i_p given in each column of Table I for $E_{C3}=+6$ was taken as unity and the other values of i_p in the same column were calculated in terms of this unit. It is seen from the data in Table II that at constant voltage on grid 1 the percentage variation in i_p as e_{g3} is varied is practically independent of the voltage on grid 1, since the difference between similarly located numbers in the columns is less than 0.05 in almost all cases. This indicates that the form of the function expressing the plate current in terms of grid-3 voltage is independent of the voltage on grid 1.

Since a change in grid-1 voltage alters the size but not the shape of the curves in Fig. 2, these curves may be expressed by the relation[4]

$$i_p = i f_3(e_{g3}) \tag{1}$$

where i is the plate current at $e_{g3}=+6$ and $f_3(e_{g3})$ is a function of grid-3 voltage determined by plotting the numbers in Table II as a function of e_{g3}. This plot is given in Fig. 3.

Now, i in (1) is a function of the voltage on grid 1. The form of this function may be obtained in a manner similar to that used in determining $f_3(e_{g3})$. In Table III are found data computed row by row from Table I, by taking the value of i_p at $E_{C1}=-3$ as unity and expressing the other values of i_p in the same row in terms of this unit. Here again, the tabular values indicate a remarkably similar variation. Except for grid-3 voltages more negative than 12 volts, the difference between similarly located numbers in the table is less than 0.03. This indicates that the functional form interrelating the plate current with the grid voltage e_{g1} is, over very wide limits, independent of the voltage on grid 3. This functional form may be determined by plotting the numbers of Table III as a function of e_{g1}. This is done in Fig. 4. Therefore, if I represents the value of i in (1) at $e_{g1}=-3$, then

$$i = I f_1(e_{g1}) \tag{2}$$

where $f_1(e_{g1})$ is the function of Fig. 4.

Substitution of (2) into (1) results in

$$i_p = I f_1(e_{g1}) f_3(e_{g3}). \tag{3}$$

This relationship has not been pointed out in previous discussions of tubes with two control grids, and forms the basis of this paper.

Equation (3) states that the plate current of a tube having two control grids acting in tandem on the same electron stream, in which the outer control grid is

[4] The list of symbols will be found in Appendix II.

TYPE 6L7G $E_P=243$ $E_{B1}=150$

TABLE I

I_P MILLIAMPERES

E_{C3}	E_{C1} −3	E_{C1} −6	E_{C1} −9	E_{C1} −12	E_{C1} −15
+22.5	15.8	9.50	5.66	3.53	2.26
+18	15.8	9.50	5.70	3.55	2.26
+12	15.6	9.43	5.68	3.55	2.27
+ 6	14.9	9.05	5.44	3.42	2.22
+ 3	14.2	8.67	5.18	3.27	2.12
0	13.5	8.19	4.92	3.13	2.02
− 2	12.4	7.59	4.54	2.88	1.90
− 4	11.2	6.87	4.11	2.64	1.75
− 6	9.9	6.06	3.67	2.37	1.57
− 8	8.5	5.19	3.15	2.04	1.36
−10	6.85	4.20	2.60	1.68	1.14
−12	5.2	3.27	2.00	1.32	0.90
−14	3.42	2.24	1.41	0.93	0.63
−16	2.02	1.39	0.88	0.60	0.41
−18	1.05	0.72	0.50	0.35	0.25
−20	0.41	0.31		0.16	0.12

TABLE II

I_P AS A FUNCTION OF E_{C3}
I_P AT $=E_{C3}=+6=$UNITY

+22.5	1.06	1.05	1.04	1.03	1.02
+18	1.06	1.05	1.05	1.04	1.02
+12	1.05	1.04	1.05	1.04	1.02
+ 6	1.00	1.00	1.00	1.00	1.00
+ 3	0.95	0.96	0.95	0.96	0.96
0	0.91	0.90	0.90	0.92	0.91
− 2	0.83	0.84	0.83	0.84	0.86
− 4	0.75	0.76	0.76	0.77	0.79
− 6	0.66	0.67	0.67	0.69	0.71
− 8	0.57	0.57	0.58	0.60	0.61
−10	0.46	0.46	0.48	0.49	0.51
−12	0.35	0.36	0.37	0.39	0.41
−14	0.23	0.25	0.26	0.27	0.28
−16	0.14	0.15	0.16	0.18	0.18
−18	0.07	0.08	0.09	0.10	0.11
−20	0.03	0.03		0.05	0.05

TABLE III

I_P AS A FUNCTION OF E_{C1}
I_P AT $E_{C1}=-3=$UNITY

E_{C1} −3	E_{C1} −6	E_{C1} −9	E_{C1} −12	E_{C1} −15
1.00	0.60	0.36	0.22	0.14
1.00	0.60	0.36	0.23	0.14
1.00	0.60	0.36	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.36	0.23	0.15
1.00	0.61	0.37	0.23	0.15
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.37	0.24	0.16
1.00	0.61	0.38	0.26	0.17
1.00	0.63	0.38	0.25	0.17
1.00	0.66	0.41	0.27	0.18
1.00	0.69	0.44	0.30	0.20
1.00	0.69	0.48	0.33	0.24
1.00	0.76		0.39	0.29

shielded on both sides, may be expressed as a constant times the product of two functions, each function involving the voltage on one control grid only. The constant is the value of the plate current at the grid voltages at which both of the functions have unity value.

Since the 6L7 and other tubes having more than one control grid have these grids as well as the screen grids interposed between the plate and the cathode, the plate is very well shielded from the cathode. In this respect the 6L7 is similar to a well-shielded pentode. In the 6L7, as in a well-shielded pentode, the plate voltage has but a minor influence on the plate current. Its chief effect is to change slightly the value of I in (3). To take account of this, we

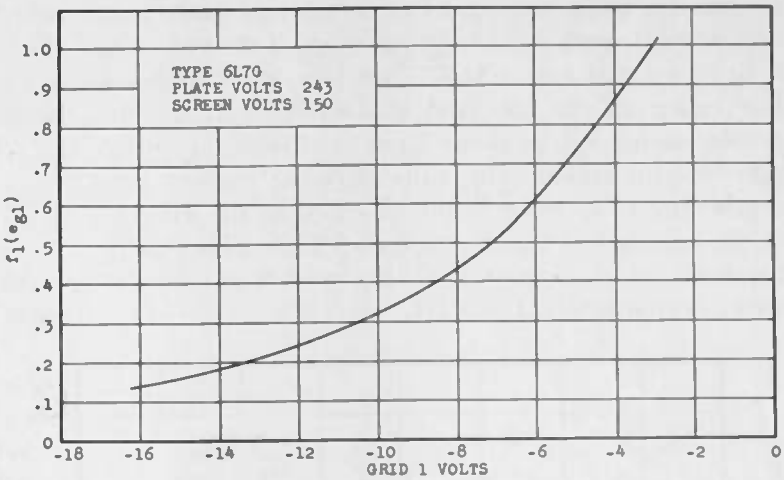

Fig. 4—Plate current as a function of grid-1 voltage expressed in terms of value at $E_{C1} = -3$ volts.

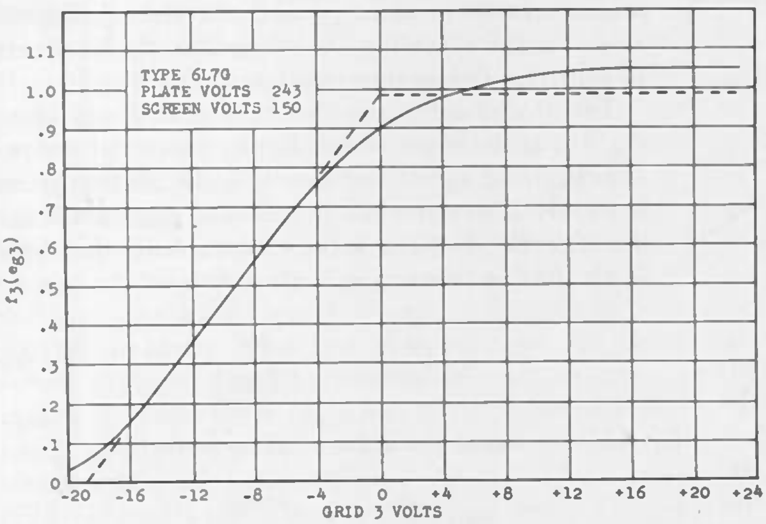

Fig. 3—Plate current as a function of grid-3 voltage expressed in terms of value at $E_{C3} = +6$ volts.

may write

$$I = I_0 f_p(e_p) \tag{4}$$

where I_0 is the value of I at the normal plate-bias voltage. For all plate voltages considerably above the screen voltage the plate current is practically independent of the plate voltage, so that $f_p(e_p)$ has a value very close to unity in this range.

The screen voltage is an independent and arbitrarily fixed parameter involved in all of the tube characteristics. Its effect on $f_3(e_{g3})$, $f_1(e_{g1})$, and $f_p(e_p)$ will now be shown.

In Fig. 5 plate current is plotted as a function of e_{g3} at a fixed voltage on grid 1, for various values of screen voltage. Here again the general form of the curves is the same as that in Figs. 2 and 3. This form is such that it may to a very close approximation be represented by three straight lines, obtained in the following manner. First, a diagonal line is drawn practically to coincide with the most linear part of the curve in the negative region, with little weight given to the roundness of the curve in the region near zero grid voltage. Second, from the intersection of this diagonal line with the axis of ordinates, a horizontal straight line is drawn extending into the region of positive grid voltage. From the intersection of the diagonal line with the axis of the abscissas, a third line is drawn along the axis of the abscissas, indicating a region in which the plate current is zero. These idealized characteristics have been indicated in Figs. 2, 3, and 5 by the dashed lines.

The idealized characteristic straight line in the positive-grid region is suggested by the flatness of the curves of Figs. 2, 3, and 5 in this region and by the fact that there is an actual saturation current which is not exceeded, no matter how positive grid 3 is driven. The plate current at grid-3 voltages causing the plate current to reach saturation (a value $e_{g3} = +22.5$ is more than sufficient) and the plate current at $e_{g3} = 0$ are not very different, the difference being about 1.2 decibels or 15 per cent. The intersection of the diagonal line with

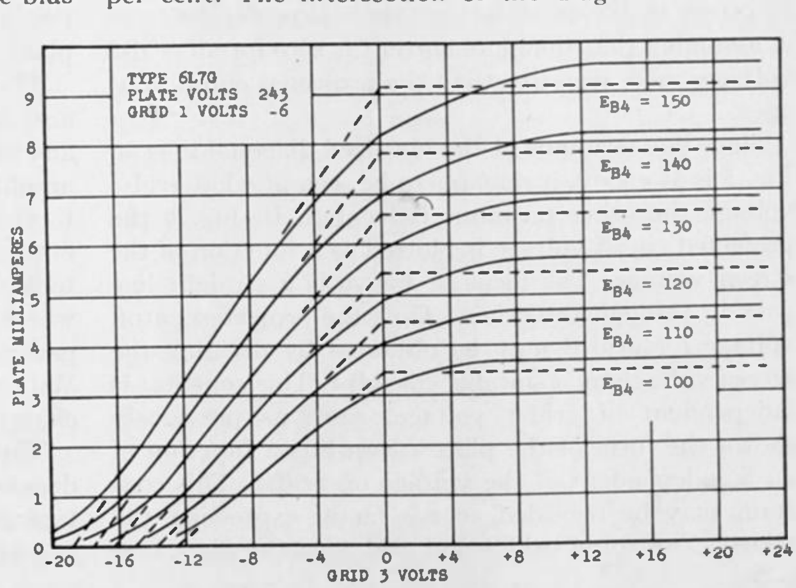

Fig. 5—Effect of grid-3 voltage on plate current at various screen voltages.

the axis of ordinates occurs at a value of plate current which is very close to the average of the values of i_p at $e_{g3}=0$ and $e_{g3}=+22.5$. Therefore, the horizontal line taken as the idealized characteristic is an approximation good to about 8 per cent over the entire positive-grid region. The value of plate current corresponding to $e_{g3}=+6$ is almost equal to the average of the currents at $e_{g3}=0$ and $e_{g3}=+22.5$. This is why the value of plate current at $E_{C3}=+6$ was taken as unity in computing Table III.

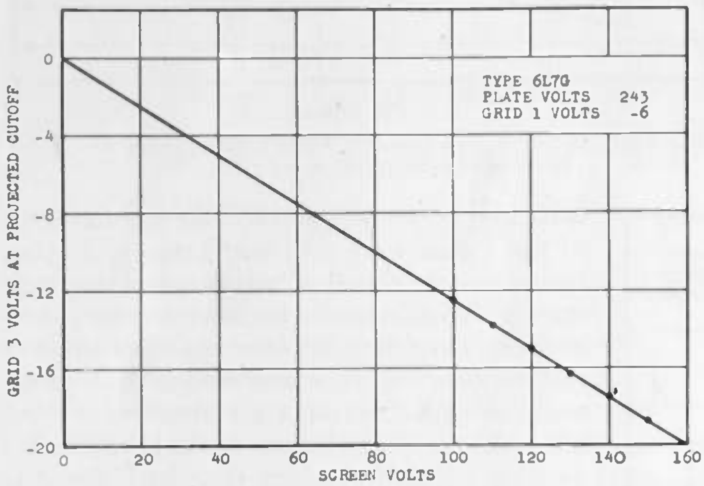

Fig. 6—Effect of screen voltage on grid-3 cutoff voltage.

The intersection of the diagonal line with the axis of the abscissas establishes a voltage which will be termed the projected cutoff voltage to distinguish it from the actual voltage required to produce actual plate-current cutoff. The actual cutoff voltage cannot be sharply defined because of the effects of irregular tube structure,[5] electron leakage around the ends of the grid structures, contact differences of potential, and initial velocity of emission of electrons from the cathode. The very small plate currents in the region near cutoff contribute practically nothing to the performance of the tube, so that there is negligible error in assuming that the plate current is zero for all grid-3 voltages more negative than the projected cutoff voltage.

Since one terminus of the idealized diagonal lines of Fig. 5 is at $e_{g3}=0$, it remains to be seen at what grid-3 voltage the other terminus is located. In Fig. 6 the projected cutoff voltage is plotted as a function of the screen voltage. The locus in Fig. 6 is a straight line passing through the origin. Thus the projected cutoff voltage for grid 3 may be obtained by dividing the screen voltage by a suitable constant. This constant is independent of grid-1 voltage, since as previously shown the form of the plate current as a function of e_{g3} is independent of the voltage on grid 1. This constant may be regarded as a μ factor expressing the relative control of grid 4 and grid 3 on the plate current. Thus, if the screen is biased to a voltage E_{B4}, and the bias required on grid 3 to produce projected cutoff is E_{3CO}, then the μ factor between grids 3 and 4 may be defined by the equation

$$\mu_{34} = -\frac{E_{B4}}{E_{3CO}}. \tag{5}$$

All voltages are taken in their algebraic or vector sense or direction with respect to the cathode.

The relationship of (5) is not even hinted at in the published data on the 6L7 tube, nor is it discussed in articles describing uses of the 6L7 tube. Without knowledge of this relationship, one can only guess as to the proper bias and control voltages for grid 3 in the many circuits to which the tube is adaptable. Equation (5) is vital in determining the form of $f_3(e_{g3})$ and therefore indispensable for the quantitative evaluation of the effects of control voltages impressed on grid 3.

The transconductance between grid 3 and plate $\partial i_p/\partial e_{g3}$ is the slope of the i_p-e_{g3} curve. In the region between $e_{g3}=0$ and $e_{g3}=E_{3CO}$ the plate current is ideally a straight-line function of e_{g3}; then if i_{PS} denotes the idealized saturation current, the slope of the line in this region is given by

$$g_{3P}\Big]_{e_{g3}=E_{3CO}}^{e_{g3}=0} = \frac{i_{PS}}{|E_{3CO}|}. \tag{6}$$

By substitution of the value of E_{3CO} from (5),

$$g_{3P}\Big]_{e_{g3}=E_{3CO}}^{e_{g3}=0} = \frac{i_{PS}\mu_{34}}{E_{B4}}. \tag{7}$$

In the region where e_{g3} is more negative than E_{3CO}, the transconductance between grid 3 and the plate is ideally zero because of plate-current cutoff, and in the region where e_{g3} is positive, the transconductance between grid 3 and the plate is ideally zero because of plate-current saturation.

The relationship between plate current and screen and grid-3 voltages as disclosed by Figs. 2, 3, 5, and 6 and by (5) is exactly the same as in an ordinary voltage-amplifier pentode. The μ factor of (5) is analogous to the cutoff amplification factor of a pentode.[6] The outer portion of the heptode is, then, virtually a pentode. There is virtually a cathode in front of grid 3 whose emission limits the plate current of the virtual pentode to the saturation currents of Figs. 2 and 5. With respect to grid 3, grids 1 and 2 act as space-charge grids fixing the emission of the virtual cathode.

The emission of the virtual cathode just described depends upon the voltages on grids 1 and 2. The nature of the control exerted by grid 1 (a control grid) at a fixed voltage on grid 2 (a screen grid) is indicated in

[5] E. L. Chaffee, "Theory of Thermionic Vacuum Tubes," McGraw-Hill Book Co., New York, N. Y., 1933, pp. 180–190.

[6] F. E. Terman, "Radio Engineering", second edition, McGraw-Hill Book Co., New York, N. Y., 1937, p. 140.

Fig. 4. This control is not linear in this tube (6L7) since grid 1 is a variable-μ grid. This variable-μ characteristic is caused primarily by the variable pitch of the grid wires, and secondarily by the fact that grids 1 and 2 are oval in shape so that the radial distances of the grid wires from the cathode are not the same in all radial directions. Further, the support rods for grid 1 have a diameter which is an appreciable fraction of even the greatest distance between grid 1 and the cathode. Thus the electrode configuration of grids 1 and 2 is that of a triode, but because of structural nonuniformity in both radial and axial dimensions, the effects are similar to what would be obtained if many smaller triodes were connected in parallel, each triode having a different μ factor[5] for grid 1. As grid 1 is made more negative, the high-μ triodes cut off their plate current first, leaving the low-μ triodes operative. Further, if the screen voltage were lowered, the high-μ triodes would cut off sooner. Therefore, there is no simple function whereby the effects of grid 1 may be described, because when either grid-1 voltage or screen voltage is changed, the effect is as though the structure of the tube were changed. However, it can be definitely stated that at lower (positive) screen voltages, the high-μ parts of the grid structure will cut off at lower (negative) grid voltages. At lower screen voltages, the plate current decreases faster as grid 1 is made more negative, so that the slope of $f_1(e_{g1})$ at a given value of e_{g1} is increased when the screen voltage is decreased.

In Tables IV and V data and computations of $f_1(e_{g1})$

TYPE 6L7G $E_P=243$ $E_{c_3}=+22.4$

TABLE IV
I_P MILLIAMPERES

E_{P_4}	E_{c_1} -3	E_{c_1} -6	E_{c_1} -9	E_{c_1} -12	E_{c_1} -16
+150	15.6	9.4	5.6	3.5	1.95
+140	13.8	8.1	4.7	2.9	1.51
+130	12.3	6.85	3.87	2.31	1.19
+120	10.4	5.46	2.99	1.72	0.88
+110	8.91	4.34	2.37	1.13	0.65
+100	7.50	3.53	1.80	0.95	0.48

TABLE V
I_P AS A FUNCTION OF E_{C_1}
I_P AT $E_{C_1}=-3=$ UNITY

+150	1.00	0.60	0.36	0.22	0.13
+140	1.00	0.59	0.34	0.21	0.11
+130	1.00	0.56	0.31	0.19	0.10
+120	1.00	0.53	0.29	0.17	0.08
+110	1.00	0.49	0.27	0.13	0.07
+100	1.00	0.47	0.24	0.13	0.06

for screen voltages between 150 and 100 volts are given. Plots of $f_1(e_{g1})$ for screen voltages of 150 and 100 are given in Fig. 7. These curves bear out the conclusions of the previous paragraph.

It has been shown above that the behavior of a tube with two control grids in tandem, having a screen grid in between the control grids and a screen grid and a suppressor grid in between the second control grid and plate, is like that of a triode in tandem with a pentode. The first grid and screen form an inner triode on whose action grid 1 has a controlling influence as in an ordinary triode. The electrons of the space current of this triode not taken up by the first screen pass through to form a virtual cathode in front of the second control grid; this virtual cathode in conjunction with the second control grid, the second screen grid, the suppressor, and the plate, form a pentode on whose action the second control grid has a controlling influence as in an ordinary pentode having an actual cathode. The plate current as controlled by the two control grids may be expressed as the product of two independent functions times a constant, the constant being the plate current corresponding to those control-grid voltages for which the respective functions have unity value. The function describing the action of the first control grid depends upon the first screen voltage and the function describing the action of the second control grid depends on the second screen voltage. When the control grid has a variable-μ structure, the control action of this grid cannot be described in terms of a μ factor. When the control grid has a constant-μ structure, then the control action of this grid can be described in terms involving this constant μ factor. In any event, at any screen voltage, $f_1(e_{g1})$ and $f_3(e_{g3})$ can be definitely determined and from these functions the behavior of the tube may be quantitatively evaluated.

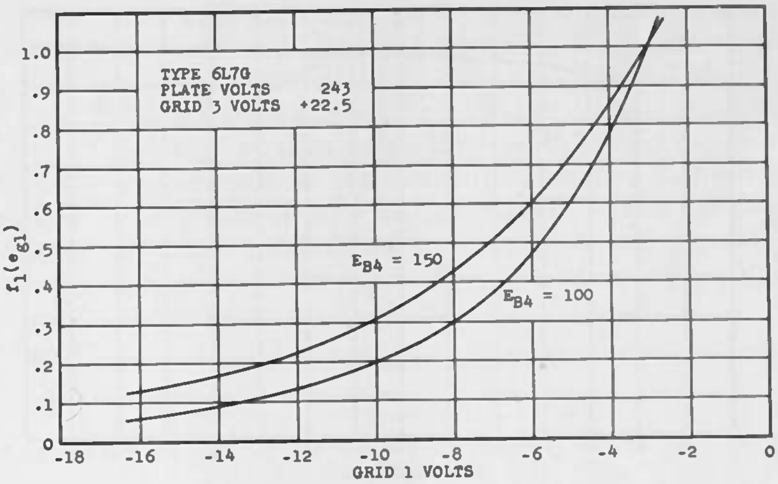

Fig. 7—Plate current as a function of grid-1 voltage at various screen voltages.

The independence of the functions $f_1(e_{g1})$ and $f_3(e_{g3})$ depends upon the electrostatic independence of the inner and outer sections of the tube. With sufficient shielding, the outer control-grid potential cannot affect the potential distribution within the screen grid so that the space current in the region between grid 2 and cathode is a function of grid-2 and grid-1 voltages only. Under these conditions the virtual emission of the virtual cathode in front of grid 3 is independent of grid-3 voltage. Therefore, (3) will apply where this independence obtains.

The form of $f_p(e_p)$ is indicated in Fig. 8, where curves of $f_p(e_p)$ for screen voltages of 100 and 150 volts

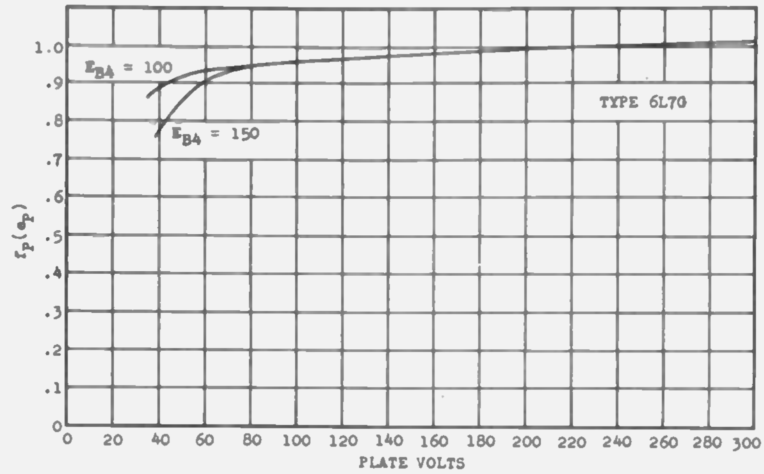

Fig. 8—Plate current as a function of plate voltage in terms of value at $E_P=240$ volts.

are plotted. The plate current at $E_P=240$ volts was taken as the unit of comparison. For plate voltages over 100 volts, the form of $f_p(e_p)$ was found to be independent of grid 1, grid 3, and screen voltages, the differences being within the accuracy (one half of 1 per cent) of the instruments used to measure the plate current. Therefore, for normal operating plate voltages, $f_p(e_p)$ is independent of grid 1, grid 3, and screen voltages.

The uses of the tube are of course not limited to those enumerated at the beginning of this paper. The tube may be applied to any communication or control purpose to which the functions $f_1(e_{g1})$ and $f_3(e_{g3})$ render it adaptable. The functions may be altered within limits by proper design of the tube structure to impart special features to $f_1(e_{g1})$ and $f_3(e_{g3})$. Since the plate current at constant screen and plate voltages is a product of two functions, the tube may be employed as a computing device. If $f_1(e_{g1})$ and $f_3(e_{g3})$ were both made linear, then the plate current could be made equal to the product of any two arbitrary functions of an independent parameter such as time, by impressing voltages which are made to vary with time in the arbitrarily desired manner.

For the quantitative handling of $f_1(e_{g1})$ and $f_3(e_{g3})$ many methods are available. The nonlinear relationships may be expressed under appropriate conditions by a power series,[7,8] by a trigonometric series,[9–11] or by an exponential series.[12] Such series expressions may be made formally precise by including an infinite number of terms. When practically applied, however, the number of terms considered is kept to a minimum consistent with the accuracy required. Even so, the calculation of the coefficients in the series representing the general relationship between plate current and grid voltage, and the calculation of another series of terms expressing the answer to a problem involving a specific grid voltage involves a great deal of labor.

In this paper the relationship between grid-3 voltage and plate current of the pentragrid heptode is on the basis of experiment given the idealized form composed of straight lines, described above and indicated in Figs. 2, 3, and 5. By an easily made classification of operating conditions, the use of infinite series is avoided in calculating the desired output frequencies when the tube is used as an amplifier, modulator, or heterodyne detector. These calculations, made from simplified formulas involving ordinary trigonometric functions and a few easily measured tube constants, require a very brief time, representing a very considerable saving over previous methods involving a multiplicity of series terms.

In the developments which follow, it has been as-

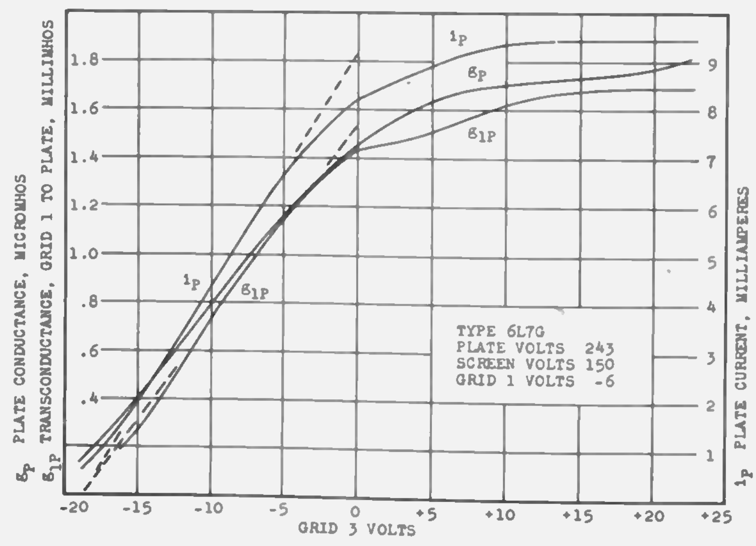

Fig. 9—Effect of grid-3 voltage on plate current, transconductance, and plate conductance.

sumed that the effects of electron transit time[13] are negligible, and that the associated circuits are such that the effects of space-charge coupling[14] are inappreciable. These conditions are readily realizable except at the very high frequencies.

[7] J. R. Carson, "A theoretical study of the three-element vacuum tube," Proc. I.R.E., vol. 7, pp. 187–200; April, 1919.

[8] E. Peterson and F. B. Llewellyn, "The operation of modulators from a physical viewpoint," Proc. I.R.E., vol. 18, pp. 38–48; January, 1930.

[9] E. Peterson and C. R. Keith, "Grid current modulation," *Bell Sys. Tech. Jour.*, vol. 7, pp. 106–139; January, 1928.

[10] J. P. Woods, "The calculation of detection performance for large signals," *Physics*, vol. 2, pp. 225–241; April, 1932.

[11] W. L. Barrow, "A contribution to the theory of nonlinear circuits with large applied voltages," Proc. I.R.E., vol. 22, pp. 964–980; August, 1934.

[12] M. J. O. Strutt, "On conversion detectors," Proc. I.R.E., vol. 22, pp. 981–1008; August, 1934.

[13] F. B. Llewellyn, "Phase angle of vacuum tube transconductance at very high frequencies," Proc. I.R.E., vol. 22, pp. 947–956; August, 1934.

[14] W. A. Harris, "The application of superheterodyne frequency conversion systems to multirange receivers," Proc. I. R. E., vol 23, pp. 279–294; April, 1935.

The complete expression for the plate current as a function of grid 1, grid 3, and plate voltages is obtained by substituting (4) into (3) with the result

$$i_p = I_0 f_1(e_{g1}) f_3(e_{g3}) f_p(e_p). \quad (8)$$

In actual service the tube may be made to operate at any of an infinite number of combinations of control-grid, screen, and plate voltages. To simplify the discussion it will be assumed for the present that grid-1, screen, and plate bias voltages are maintained constant. Then from (8), the transconductance between grid 1 and the plate

$$g_{1P} = \frac{\partial i_p}{\partial e_{g1}} = I_0 \frac{df_1(e_{g1})}{de_{g1}} f_3(e_{g3}) f_p(e_p) \quad (9)$$

and the plate conductance,

$$g_P = \frac{\partial i_p}{\partial e_p} = I_0 f_1(e_{g1}) f_3(e_{g3}) \frac{df_p(e_p)}{de_p} \quad (10)$$

should both vary in the same manner when the voltage on grid 3 is varied. In Fig. 9 these quantities are plotted for the conditions of fixed grid-1, screen, and plate bias voltages as indicated. All the curves have the same shape. The curves of i_p and g_{1P} have the same projected cutoff as that given for $f_3(e_{g3})$ in Fig. 3. The projected cutoff of plate conductance is not quite the same. Even with perfect insulation, the measured value of plate conductance would never be zero because of ionization of the residual gas in the tube. But the contribution of the electron stream to the plate conductance is zero when the voltage on grid 3 is sufficiently negative to prevent cathode electrons from reaching the plate. Since the plate resistance is high in the region near cutoff, very little error will be caused by assuming that there is a projected cutoff of plate conductance at the same point as there is a projected cutoff of thermionic plate current. Then the plate current, transconductance from grid 1 to plate, and plate conductance are varied in the manner of $f_3(e_{g3})$ when e_{g3} is varied.

The idealized values of i_p, g_{1P}, and g_P in the positive region of grid-3 voltage will be called the saturation values. The values used in computations for the saturation values will be the average of measurements made on the tube at $E_{C3}=0$ and $E_{C3}=+22.5$. These values will be denoted by capital letters, namely I_{PS}, G_{1P}, and G_{PS}. Further, if these values had been used in the computation of Table II, the value of $f_3(e_{g3})$ at $E_{C3}=0$ would have been closer to unity than it is in Fig. 3. Therefore, $f_3(e_{g3})$ will be assumed equal to unity for all positive values of e_{g3}, decreasing linearly to zero at $e_{g3}=E_{3CO}$ as e_{g3} is made more negative. This idealized form is indicated in Fig. 10. Therefore, it is evident that the three measured quantities needed to specify the behavior of the pentagrid heptode are the saturation values of plate current, grid-1-to-plate transconductance, and plate conductance, denoted respectively by I_{PS}, G_{1P}, and G_{PS}. These and the μ factor μ_{34}, which is independent of electrode voltages, are all that are required to calculate the performance of the pentagrid heptode by the methods hereinafter outlined.

At constant bias voltages on grid 1, screen, and plate, the plate current, transconductance, and plate conductance of (8), (9), and (10), respectively, may be expressed as

$$i_p = I_{PS} f_3(e_{g3}) \quad (11)$$

$$g_{1P} = G_{1P} f_3(e_{g3}) \quad (12)$$

$$g_P = G_{PS} f_3(e_{g3}) \quad (13)$$

and from (13) the plate resistance may be expressed as

$$r_P = \frac{1}{g_P} = \frac{1}{G_{PS} f_3(e_{g3})} = \frac{R_{PS}}{f_3(e_{g3})} \quad (14)$$

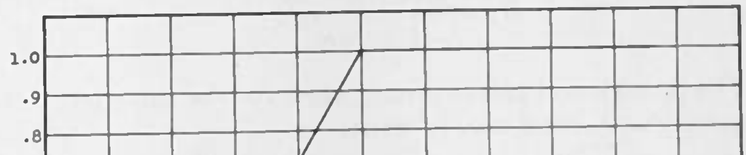

Fig. 10—Idealized form of $f_3(e_{g3})$.

where R_{PS} is the saturation value of the plate resistance.

The Pentagrid Heptode as an Amplifier

The equivalent circuit of a tube acting as a class A1 amplifier may be regarded as that of a constant-voltage generator feeding a circuit consisting of the plate resistance of the tube in series with the plate-load impedance, or alternatively, the equivalent circuit may be regarded as that of a constant-current generator feeding a circuit consisting of the plate resistance and load impedance in parallel.[6] The constant-voltage equivalent circuit has the same form as equivalent circuits which may be set up on the basis of Thevenin's theorem, and the constant-current equivalent circuit has the same form as equivalent circuits which may be set up by applying a theorem attributed to

[6] *Loc. cit.*, p. 172.

Norton.[15] When grid 3 of the pentagrid heptode is held at a constant voltage E_{C3}, the tube may be made to serve as an amplifier having a fixed transconductance. The only tube elements having incremental alternating voltages will be grid 1 and the plate, so that

$$di_p = \frac{\partial i_p}{\partial e_{g1}} de_{g1} + \frac{\partial i_p}{\partial e_p} de_p \tag{15}$$

$$= g_{1P} de_{g1} + g_P de_p \tag{16}$$

plus negligible higher-order terms. For a load resistance R_L

$$de_p = -R_L di_p \tag{17}$$

so that

$$di_p = \frac{g_{1P} de_{g1}}{1 + g_P R_L} \cdot \tag{18}$$

When

$$de_{g1} = E_{S1} \cos \omega_1 t \tag{19}$$

then

$$di_p = \frac{g_{1P} E_{S1} \cos \omega_1 t}{1 + g_P R_L} \cdot \tag{20}$$

The incremental plate current will have the same frequency as E_{S1} and may be written as

$$di_p = I_P \cos \omega_1 t$$

where

$$I_P = \frac{g_{1P} E_{S1}}{1 + g_P R_L} \tag{21}$$

$$= g_{1P} E_{S1} \frac{r_P}{r_P + R_L} \cdot \tag{22}$$

At the operating bias voltage E_{C3} and after the substitution of values from (12), (13), and (14), (22) becomes

$$I_P = G_{1P} f_3(E_{C3}) \frac{\dfrac{R_{PS}}{f_3(E_{C3})}}{\dfrac{R_{PS}}{f_3(E_{C3})} + R_L} E_{S1}. \tag{23}$$

Equation (23) also holds in complex form for a complex load impedance $Z_L = R_L + jX_L$, when Z_L is substituted for R. Thus,

$$I_P = G_{1P} f_3(E_{C3}) \frac{\dfrac{R_{PS}}{f_3(E_{C3})}}{\dfrac{R_{PS}}{f_3(E_{C3})} + Z_L} E_{S1}. \tag{24}$$

Equation (24) is the equation for the constant-current form of the equivalent circuit. Thus the tube acts as a generator, of constant current $G_{1P} f_3(E_{C3}) E_{S1}$, feeding a load consisting of the external load impedance Z_L in parallel with the internal plate resistance $R_{PS}/f_3(E_{C3})$. It is obvious from (24) and Fig. 10 that the gain may be made to increase and decrease if the magnitude of $f_3(E_{C3})$ were made to increase and decrease. The fact that the gain may be varied by varying E_{C3} adapts the pentagrid heptode for use as a variable-gain amplifier for volume expansion, volume contraction, or volume limiting at audio or radio frequencies. Some circuits in which the 6L7 is so used have been described elsewhere.[16–18] The gain does not vary linearly with $f_3(E_{C3})$, however, since as the transconductance increases, the plate resistance decreases.

The Pentagrid Heptode as a Modulator

The pentagrid heptode may also be used as a double-control-grid modulator. For the most linear modulation, the carrier voltage should be applied to grid 1, the modulating voltage should be applied to grid 3, and the voltage on grid 3 held within the range

$$E_{3CO} < e_{g3} < 0. \tag{25}$$

With the voltage on grid 3 consisting of a bias voltage E_{C3} and an oscillator voltage $E_{O3} \cos \omega_3 t$,

$$e_{g3} = E_{C3} + E_{O3} \cos \omega_3 t, \tag{26}$$

the values of $f_3(e_{g3})$ vary cyclically with time. The function $f_3(e_{g3})$ may then be expressed as a Fourier series in terms of the parameter $\omega_3 t$. Then if (25) holds,

$$f_3(e_{g3}) = a + b \cos \omega_3 t \tag{27}$$

there being no higher frequency terms. The values of a and b are derived in Appendix I, Case 1. When the plate load impedance is resistive to carrier and sideband frequencies, and zero at other frequencies, (21) holds but g_{1P} and g_P then vary cyclically with time in accordance with (27), (12), and (13). Then the carrier and modulation terms of the plate current are given by

$$di_p = \frac{G_{1P}(a + b \cos \omega_3 t) E_{S1} \cos \omega_1 t}{1 + R_L G_{PS}(a + b \cos \omega_3 t)} \tag{28}$$

$$\doteq E_{S1} G_{1P}(a + b \cos \omega_3 t) \cos \omega_1 t \cdot \left\{ \frac{1}{1 + R_L G_{PS}(a + b \cos \omega_3 t)} \right\}. \tag{29}$$

Equation (29) may be written in the form

[15] W. L. Everitt, "Communication Engineering," second edition, McGraw-Hill Book Co., New York, N. Y., 1937, pp. 47–49.

[16] Application Note No. 53, The 6L7 as a volume expander for phonographs, RCA Manufacturing Co., Inc., November 27, 1935.

[17] Application Note No. 57, The 6L7 as an R-F amplifier, RCA Manufacturing Co., Inc., February 5, 1936.

[18] J. J. Lamb, "A noise-silencing I.F. circuit for superheterodyne receivers, *QST*, vol. 20, pp. 11–14 *et seq.*; February, 1936.

$$di_p = E_{S1}G_{1P}(a + b \cos \omega_3 t) \cos \omega_1 t \cdot \left\{\frac{1}{1 + aR_L G_{PS}}\right\}\left\{\frac{1}{1 + q \cos \omega_3 t}\right\} \quad (30)$$

where

$$q = \frac{bR_L G_{PS}}{1 + aR_L G_{PS}}. \quad (31)$$

If

$$q \ll 1$$

or, what is equivalent, if

$$bR_L G_{PS} \ll 1 + aR_L G_{PS}. \quad (32)$$

then the last term of (30) would remain practically equal to unity, so that

$$di_p = \frac{E_{S1}aG_{1P}\left(1 + \frac{b}{a} \cos \omega_3 t\right) \cos \omega_1 t}{1 + R_L \frac{a}{R_{PS}}}. \quad (33)$$

The incremental plate current of carrier and sideband frequencies may be written as

$$di_p = I_P(1 + m \cos \omega_3 t) \cos \omega_1 t$$

where

$$m = \frac{b}{a} \quad (34)$$

so that (32) becomes

$$I_P(1+m \cos \omega_3 t) \cos \omega_1 t = E_{S1}aG_{1P}\left(1+ \frac{b}{a} \cos \omega_3 t\right) \cos \omega_1 t \left\{\frac{\frac{R_{PS}}{a}}{\frac{R_{PS}}{a} + R_L}\right\}. \quad (35)$$

Thus the equivalent circuit of the tube acting as an amplitude modulator is that of a tube having a transconductance aG_{1P} and an internal plate resistance R_{PS}/a, feeding a plate load R_L, and having its signal $E_{s1} \cos \omega_1 t$ modulated to a degree $m = b/a$. The condition of (32) is essential if distortion is to be avoided.

The Pentagrid Heptode as a Heterodyne Detector

In its use as a heterodyne detector, the pentagrid heptode is essentially a modulator with the received signal applied to grid 1 and a voltage derived from a local oscillator applied to grid 3. The plate load impedance is made responsive to the difference frequency only. Since such a circuit selects but one component resulting from the modulation process, the object of modulation ceases to be that of maximum output consistent with the required linearity. Linearity with respect to the oscillator voltage is no longer required. The object is to produce as large a difference-frequency component as possible. To this end, the oscillator voltage applied to the tube is made as large as possible. Further, it is not necessary to restrict the values of grid-3 voltage to the linear region of $f_3(e_{g3})$ between zero and cutoff. Under these conditions, the Fourier series for $f_3(e_{g3})$ in terms of the parameter $\omega_3 t$ will contain terms which are harmonic frequencies of the fundamental oscillator frequency $\omega_3/2\pi$. Since the fundamental produces the desired difference-frequency output, the harmonic terms will be omitted in the equations which follow.

Since

$$e_{g3} = E_{C3} + E_{O3} \cos \omega_3 t \quad (36)$$

therefore

$$f_3(e_{g3}) = a + b \cos \omega_3 t + \cdots \quad (37)$$

Then from (12), (13), and (14),

$$g_{1P} = G_{1P}(a + b \cos \omega_3 t) \quad (38)$$

$$g_P = G_{PS}(a + b \cos \omega_3 t) \quad (39)$$

$$r_P = \frac{R_{PS}}{a + b \cos \omega_3 t}. \quad (40)$$

Equation (16) now becomes

$$di_p = G_{1P}(a + b \cos \omega_3 t)de_{g1} + G_{PS}(a + b \cos \omega_3 t)de_p \quad (41)$$

$$= G_{1P}(a + b \cos \omega_3 t)E_{S1} \cos \omega_1 t + G_{PS}(a + b \cos \omega_3 t)de_p. \quad (42)$$

When the plate load impedance is zero, de_p is zero, and (42) may be used to evaluate the conversion transconductance. Since in converter service the oscillator frequency is usually higher than the signal frequency, the difference-frequency current will be taken as that having the frequency $(\omega_3 - \omega_1)/2\pi$. Then, with $de_p = 0$, expansion of (42) results in

$$di_p = E_{S1}G_{1P}a \cos \omega_1 t + E_{S1}G_{1P}\frac{b}{2} \cos (\omega_3 + \omega_1)t + E_{S1}G_{1P}\frac{b}{2} \cos (\omega_3 - \omega_1)t. \quad (43)$$

With the difference-frequency component denoted by

$$(di_p)\omega_3 - \omega_1 = I_d \cos (\omega_3 - \omega_1)t \quad (44)$$

this may be set equal to the last term of (43) so that

$$I_d \cos (\omega_3 - \omega_1)t = E_{S1}G_{1P}\frac{b}{2} \cos (\omega_3 - \omega_1)t. \quad (45)$$

The conversion transconductance between grid 1 and the plate is the ratio of I_d to E_{S1}, so that when the conversion transconductance is denoted by the symbol S_{1P} then

$$S_{1P}=G_{1P}\frac{b}{2}\cdot \tag{46}$$

The factor b of (45) is the value of the amplitude of the fundamental component of a wave obtained by plotting values of $f_3(e_{g3})$ against values of the parameter $\omega_3 t$. Such a wave is a pulsating unidirectional wave, with maximum value of unity. The greatest possible value of b for such a wave is $2/\pi$ for a wave shape having pulses rectangular in shape and one-half cycle long. Therefore b in (45) can never exceed $2/\pi$ and the conversion transconductance can never exceed $1/\pi$ times the maximum possible value of G_{1P}.

The voltage applied to grid 3 has the form indicated by (36); where E_{C3} is the steady bias voltage and E_{O3} is the amplitude (peak value) of the oscillator driving voltage. In the Fourier analysis, the symmetry of the pulses in the value of $f_3(e_{g3})$ renders sufficient an integration extending over only the first half of the first cycle of the oscillator voltage appearing in (36). When on the positive swing of the oscillator voltage, grid 3 is driven positive, the value of $f_3(e_{g3})$ will ideally be constant at the saturation value of unity (see Fig. 10) while grid 3 is positive, or during that angular fraction of the cycle between $\omega_3 t=0$ and $\omega_3 t=\phi_S$ where ϕ_S is obtained by setting (36) equal to zero. Whence

$$\cos\phi_S=-\frac{E_{C3}}{E_{O3}}\cdot \tag{47}$$

When, on the negative swing of the oscillator voltage, grid 3 is driven more negative than projected cutoff voltage, the value of $f_3(e_{g3})$ ideally will be zero during that angular fraction of the cycle between $\omega_3 t=\phi_0$ and $\omega_3 t=\pi$, where ϕ_0 is obtained by setting (36) equal to E_{3CO}. Whence

$$\cos\phi_0=\frac{E_{3CO}-E_{C3}}{E_{O3}}=\frac{-\dfrac{E_{B4}}{\mu_{34}}-E_{C3}}{E_{O3}}\cdot \tag{48}$$

The oscillator and bias voltages applied to grid 3 may be such as to cause neither cutoff nor saturation, either cutoff or saturation, or both cutoff and saturation. The following classification of operating conditions covers all cases where the bias voltage on grid 3 is negative.

Case I. No Cutoff, No Saturation

In this case, grid 3 is not biased nor driven more negative than projected cutoff voltage, nor is it driven positive with respect to the cathode. Operation is confined to the region of the idealized diagonal line of Fig. 10.

Case II. Cutoff, No Saturation

In this case, grid 3 is biased or is driven more negative than projected cutoff voltage, but is not driven positive. Operation extends over the region of the idealized diagonal line of Fig. 10 and into the region where $f_3(e_{g3})$ is zero.

Case III. Saturation, No Cutoff

In this case grid 3 is driven positive with respect to the cathode, but is not biased or driven more negative than projected cutoff voltage. Operation extends over the region of the idealized diagonal lines and into the region of the idealized horizontal line to the left of the axis of ordinates in Fig. 10.

Case IV. Saturation and Cutoff

In this case, grid 3 is driven positive and also is driven or is biased so negative as to cause plate-current cutoff. Operation extends over the entire region of the diagonal line of Fig. 10 and into the regions on both sides.

The factor b in (46) can be determined by Fourier analysis by evaluating integrals suitable to the regions over which the tube operates. The idealized characteristic composed of straight lines makes these integrals very simple. The values for b for the four cases are derived in Appendix I. The complete formulas for the conversion transconductance are given below.

Case I. No Cutoff, No Saturation

$$S_{1P}=E_{O3}\frac{G_{1P}\mu_{34}}{2E_{B4}}\cdot \tag{49}$$

Case II. Cutoff, No Saturation

$$S_{1P}=E_{O3}\frac{G_{1P}\mu_{34}}{2E_{B4}}\left\{\frac{\phi_0-\sin\phi_0\cos\phi_0}{\pi}\right\}\cdot \tag{50}$$

Case III. Saturation, No Cutoff

$$S_{1P}=E_{O3}\frac{G_{1P}\mu_{34}}{2E_{B4}}\left\{1-\frac{\phi_S-\sin\phi_S\cos\phi_S}{\pi}\right\}\cdot \tag{51}$$

Case IV. Saturation and Cutoff

$$S_{1P}=E_{O3}\frac{G_{1P}\mu_{34}}{2E_{B4}}\left\{\frac{\phi_0-\sin\phi_0\cos\phi_0}{\pi}-\frac{\phi_S-\sin\phi_S\cos\phi_S}{\pi}\right\}\cdot \tag{52}$$

Case I represents a condition mathematically similar to that for square-law detection, since the difference-frequency current, which may be obtained by multiplying (49) by E_{S1}, is directly proportional to the product of the voltage amplitudes $E_{S1}E_{3O}$.

For cases II and IV, the cutoff angle ϕ_0 is given by (48) which may be restated as follows, the bars around E_{C3} indicating magnitude only:

$$\cos\phi_0=\frac{|E_{C3}|-\dfrac{E_{B4}}{\mu_{34}}}{E_{O3}}\cdot \tag{53}$$

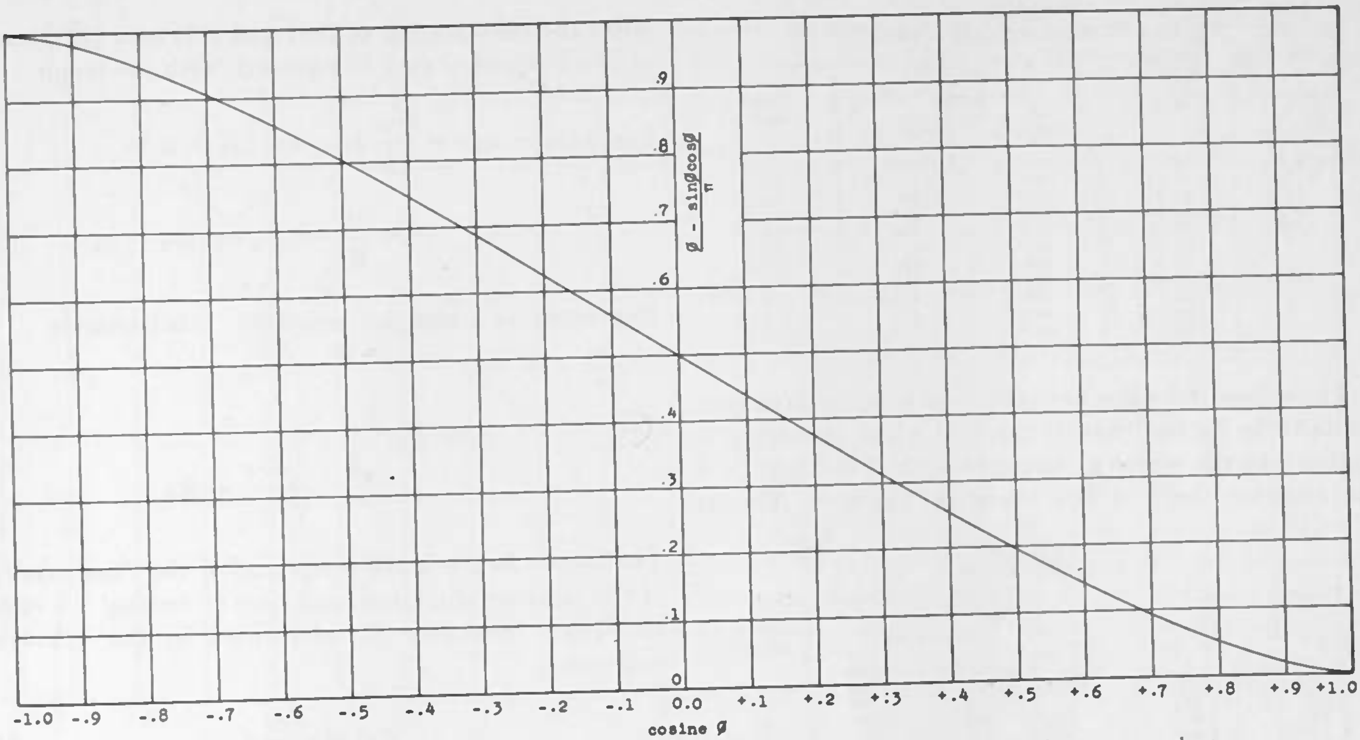

Fig. 11—Curve for value of trigonometric expression occurring in the formulas for conversion transconductance.

For Cases III and IV, the saturation angle ϕ_S is given by (9) which, restated in terms of magnitudes only is

$$\cos \phi_S = \frac{|E_{C3}|}{E_{O3}}. \tag{54}$$

Cos ϕ_0 is a positive number only when the magnitude of the negative bias on grid 3 is greater than that required for projected cutoff. Cos ϕ_S is always a positive number. The values of the trigonometric expressions in (50), (51), and (52) may be obtained from Fig. 11.

In the tables which follow, the conversion conductance calculated by the method outlined above is compared with the measured value, and the error is given to the nearest 0.1 decibel. For Tables VI and IX inclusive, the tube is a type 6L7G; the values given are for conditions where $E_{C1}=-6.0$ volts, $E_{B4}=150$ volts, $E_P=243$ volts, $G_{1P}=1560$ micromhos, $\mu_{34}=8.0$; the bias voltages on grid 3 were obtained from a battery source and were independent of the oscillator voltages applied to grid 3.

TABLE VI
CASE I. NO SATURATION, NO CUTOFF

E_{C3}	E_{O3}	S_{1P} (micromhos) calculated	measured	Error (decibels)
− 6.2	3.17	132	127	+0.3
− 7.7	"	"	136	−0.2
− 8.9	"	"	143	−0.7
−10.5	"	"	149	−1.0
−12.3	"	"	148	−1.0
−13.8	"	"	138	−0.4
− 8.9	1.58	66	77	−1.3
"	3.17	132	143	−0.7
"	4.55	189	204	−0.7
"	6.16	256	266	−0.3
"	7.56	315	308	+0.2
"	9.00	374	350	+0.6

When grid-3 bias is secured by means of a grid resistor, the grid is always driven positive on the positive swing of the oscillator voltage. The mode of operation is that of Case III for small oscillator voltages and Case IV for large oscillator voltages. Table X presents data for the operation of the same tube as above, with grid-3 bias secured by means of a grid resistor of 51,600 ohms. The values of grid-3 bias were determined by measuring the average grid current. The operating voltages on the other electrodes were the same as those for Tables VI to IX.

TABLE VII
CASE II. CUTOFF, NO SATURATION

E_{C3}	E_{O3}	S_{1P} (micromhos) calculated	measured	Error (decibels)
−12.3	7.56	304	294	+0.3
"	9.00	342	334	+0.2
"	9.50	352	346	+0.1
"	11.04	390	379	+0.2
"	11.82	410	392	+0.4
"	13.21	439	416	+0.4
−16.7	3.17	116	99	+1.4
"	6.16	181	173	+0.4
"	9.00	241	242	0.0
"	9.50	251	256	−0.2
"	11.82	299	301	−0.1
"	15.22	370	366	+0.1
"	16.60	399	388	+0.2

TABLE VIII
CASE III. SATURATION, NO CUTOFF

E_{C3}	E_{O3}	S_{1P} (micromhos) calculated	measured	Error (decibels)
− 6.2	7.56	301	278	+0.7
"	9.00	337	327	+0.3
"	9.50	350	338	+0.3
"	11.82	403	393	+0.2
"	13.21	432	421	+0.2
− 7.7	8.72	355	338	+0.4
"	9.50	376	358	+0.4
"	10.27	397	384	+0.3
0.0	4.06	85	94	−0.9
"	7.95	166	175	−0.5
"	11.82	248	272	−0.8
"	15.22	317	358	−1.1
"	18.61	387	426	−0.8

TABLE IX
CASE IV. SATURATION AND CUTOFF

E_{C3}	E_{O3}	S_{1P} (micromhos) calculated	measured	Error (decibels)
−6.2	15.22	448	440	+0.2
"	18.61	466	483	−0.3
"	22.4	476	500	−0.4
−10.5	11.82	433	407	+0.5
"	15.22	460	440	+0.4
"	18.61	473	466	+0.1
"	22.4	480	489	−0.2
−16.8	18.61	421	411	+0.2
"	22.4	450	431	+0.4
"	30.0	468	483	−0.3
"	38.2	482	512	−0.5

The calculated value of conversion transconductance tends to be higher than the actual when operation is confined to the region around zero grid-3 voltage, as is the case for the first few items in Table X. This is because the idealized form of $f_3(e_{g3})$ is given a sharper bend at zero grid-3 voltage than is possessed by the actual characteristic, as shown in Fig. 3. At the very large oscillator voltages, the calculated value of conversion transconductance tends to be less than the measured value, because the large oscillator voltages drive grid 3 far into the positive-grid region, where the idealized form of $f_3(e_{g3})$ is about 0.6 decibel less than the actual. For most cases the calculated and measured values agree within the accuracy of measurement, which is estimated to be 0.5 decibel.

TABLE X
CASES III AND IV

E_{O3}	E_{C3}	Case	S_{1P} (micromhos) calculated	measured	Error (decibels)
3.17	−2.32	III	121	96	+2.0
4.55	−3.35	"	174	149	+1.3
6.16	−4.49	"	235	211	+0.9
9.00	−6.45	"	342	327	+0.4
11.82	−8.30	IV	434	411	+0.5
15.22	−10.83	"	460	441	+0.4
22.4	−15.11	"	462	461	0.0
30.0	−19.76	"	456	466	−0.2
36.8	−22.6	"	458	472	−0.3

When the external plate impedance of the converter tube is not large compared to the internal plate resistance, the conversion gain may be calculated with sufficient accuracy by simply multiplying the conversion transconductance by the external impedance in the plate circuit. The error in this simple method is less than 1 decibel when the external plate impedance is less than one eighth of the internal plate resistance. In the pentagrid converter tube the plate is shielded from the cathode by 5 grids. The plate resistance, therefore, is very high. However, the external plate impedance offered by the first intermediate-frequency transformer in a frequency-conversion system may also be very high, so that it may be necessary to take the plate resistance into account in calculating the conversion gain. When the external plate load impedance has resistive impedance R_d to currents of the difference frequency and zero impedance to currents of other frequencies, then

$$de_p = -R_d I_d \cos(\omega_3 - \omega_1)t. \tag{55}$$

After the substitution of (55) and (44) into (42), terms of like frequency may be equated, with the result

$$I_d \cos(\omega_3 - \omega_1)t = \frac{G_{1P}}{2} b E_{S1} \cos(\omega_3 - \omega_1)t - \frac{a}{R_{PS}} R_d I_d \cos(\omega_3 - \omega_1)t. \tag{56}$$

Expressed as a complex equation, this becomes

$$I_d = E_{S1}\frac{G_{1P}}{2} b \frac{\frac{R_{PS}}{a}}{\frac{R_{PS}}{a} + R_d}. \tag{57}$$

The factor R_{PS}/a is the reciprocal of the time average of the plate conductance, and may be termed the effective plate resistance R_p' as defined by the following equation:

$$R_P' = \frac{R_{PS}}{a}. \tag{58}$$

A similar derivation for the case of a complex load impedance having a value Z_d at the difference frequency and zero impedance at other frequencies would yield a more general form of (57), which, after substitutions according to (46) and (58), would be

$$I_d = E_{S1} S_{1P} \frac{R_P'}{R_P' + Z_d}. \tag{59}$$

This is the equation for the difference-frequency current I_d in the load impedance Z_d which would exist in an equivalent circuit consisting of a constant-current generator generating a current of amplitude $E_{S1}S_{1P}$ *at the difference frequency* feeding a load composed of a resistance equal to the effective plate resistance R_P' connected in parallel with the difference-frequency impedance Z_d. The difference-frequency voltage across the difference-frequency plate load impedance is I_dZ_d, so that the conversion gain is given by the equation

$$\text{conversion gain} = S_{1P} \frac{R_P' Z_d}{R_P' + Z_d}. \tag{60}$$

The equivalent resistance R_P' is the reciprocal of the time average of the plate conductance under converter operating conditions with both bias and oscillator driving voltages applied to grid 3. It is not the same as the value that would be measured if the oscillator voltage were removed but the same bias voltage retained. For example, under actual converter conditions, grid 3 might be biased beyond cutoff. The measured value of plate resistance at this bias voltage ideally would be infinite, since the electron stream would not reach the plate. But in converter service at this bias the time

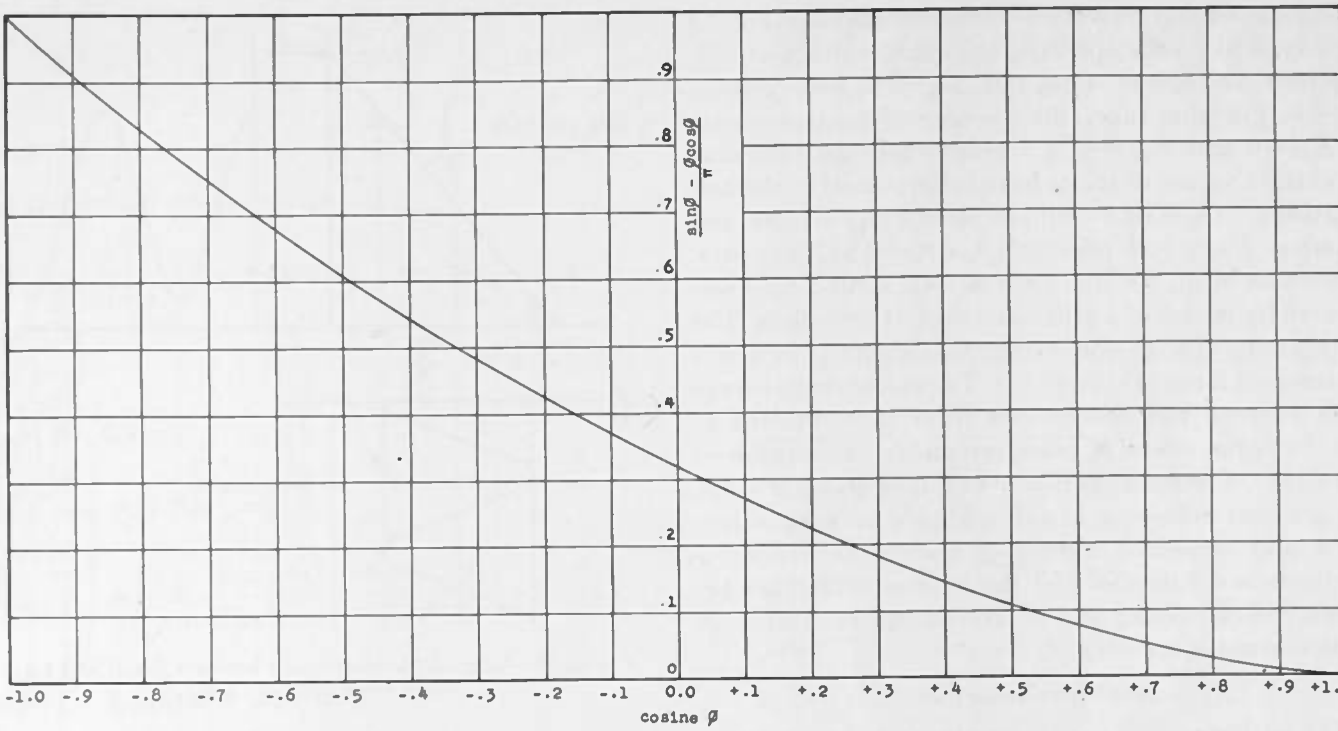

Fig. 12—Curve for value of trigonometric expression occurring in the formulas for factor a.

average of the plate current would not be zero; therefore, the time average of the plate conductance could not be zero, and the time average of the plate resistance would not be infinite. The value of R_P' is ideally that value which would be measured if E_{C3} were adjusted to make the plate current equal to the time average value under actual converter conditions.

The time average of the plate current under converter conditions may be calculated as follows: with all electrode voltages constant except those on grid 3, to which an oscillator voltage $E_{O3}\cos\omega_3 t$ as well as a bias voltage E_{C3} is applied, the plate current may be represented by (11), which after substitution from (37), becomes

$$i_p = I_{PS}(a + b\cos\omega_3 t). \tag{61}$$

If e_{g1} were now varied by the signal voltage, an incremental current would flow whose value is given by (41). But if the plate load impedance were zero to all but difference-frequency currents, de_p in (41) would be given by (55). Then all the terms in (41) would be alternating terms with average values of zero. Then the average plate current I_{Pa} would be given by the constant term of (61), so that

$$I_{Pa} = aI_{PS}. \tag{62}$$

The values of a are derived in Appendix I. The formulas for the four classifications of operating conditions are given below.

Case I. No Cutoff, No Saturation

$$a = 1 - \frac{\mu_{34}}{E_{B4}}\left| E_{C3} \right|. \tag{63}$$

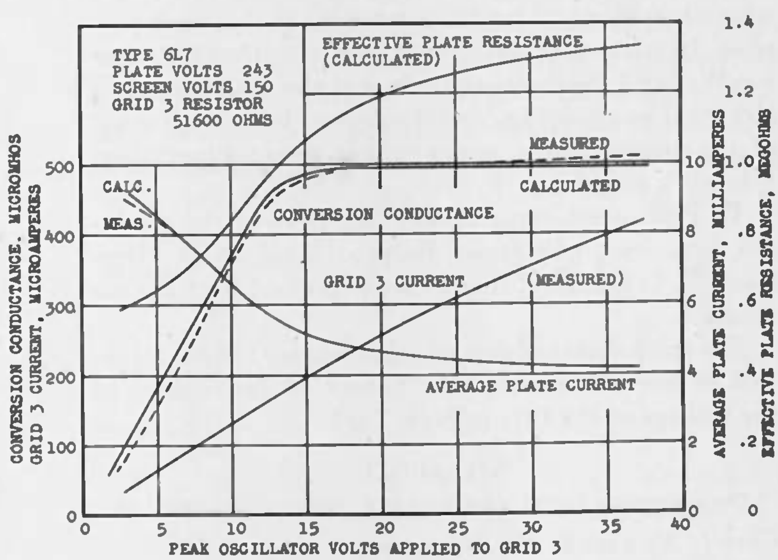

Fig. 13—The pentagrid heptode as a heterodyne detector.

Case II. Cutoff, No Saturation

$$a = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\sin\phi_0 - \phi_0\cos\phi_0}{\pi}\right\}. \tag{64}$$

Case III. Saturation, No Cutoff

$$a = 1 - E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\cos\phi_S + \frac{\sin\phi_S - \phi_S\cos\phi_S}{\pi}\right\}. \tag{65}$$

Case IV. Saturation and Cutoff

$$a = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\sin\phi_0 - \phi_0\cos\phi_0}{\pi} - \frac{\sin\phi_S - \phi_S\cos\phi_S}{\pi}\right\}. \tag{66}$$

The values of the trigonometric expressions in (63) to (66) may be obtained from Fig. 12.

Calculated and measured results are given in Fig. 13 for a type 6L7 tube operated at a plate voltage of 243, a screen voltage of 150, and a grid-1 bias voltage of -6. For this tube, the average of measurements at $E_{C3}=0$ and $E_{C3}=+22.5$ established the following saturation values of the tube constants used in the calculations: $I_{PS}=10.3$ milliamperes, $G_{1P}=1660$ micromhos, $G_{PS}=1.88$ micromhos, $R_{PS}=0.532$ megohm. The value of μ_{34} for this tube is 7.42. Grid-3 bias was secured by means of a grid resistor of 51,600 ohms. The calculated value of conversion transconductance was determined from (51) and (52). The calculated average plate current was determined from (62). Plotted in Fig. 13 is the effective plate resistance as determined from (58). For peak oscillator voltages above 9 volts, the greatest difference in this instance between calculated and measured values of conversion transconductance is 0.3 decibel and the greatest difference between the calculated and measured values of average plate current is 0.1 decibel.

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Appendix I

Derivation of Wave-Form Constants a and b

Case I. No Cutoff, No Saturation

The operation of the tube is confined to the region of linear variation of $f_3(e_{g3})$. Then

$$f_3(e_{g3}) = \frac{1}{|E_{3co}|}(|E_{3co}| - |e_{g3}|).$$

The first coefficient of the Fourier series is given by

$$a = \frac{1}{2\pi}\int_0^{2\pi} f_3 d\phi = \frac{1}{\pi}\int_0^{\pi} f_3 d\phi.$$

The second coefficient of the Fourier series is given by

$$b = \frac{1}{\pi}\int_0^{2\pi} f_3 \cos\phi d\phi = \frac{2}{\pi}\int_0^{\pi} f_3 \cos\phi d\phi.$$

From the above,

$$a = 1 - \frac{\mu_{34}}{E_{B4}}|E_{C3}|$$

$$b = E_{O3}\frac{\mu_{34}}{E_{B4}}\cdot$$

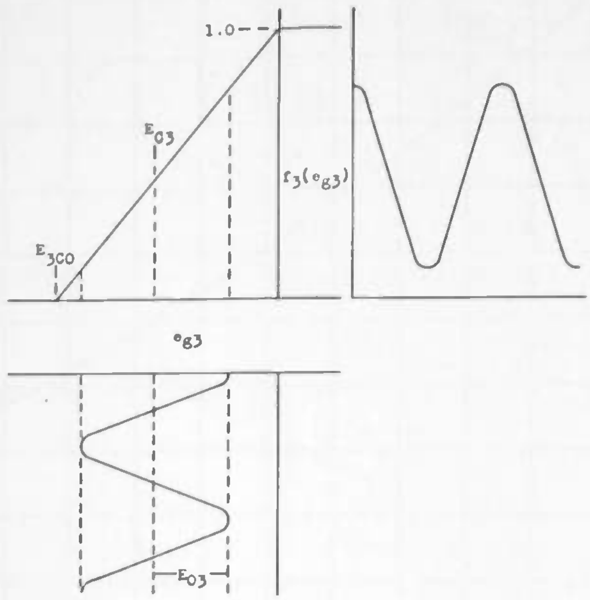

Fig. 14—Wave-form relationship between $f_3(e_{g3})$ and e_{g3} under conditions of Case I.

Case II. Cutoff, No Saturation

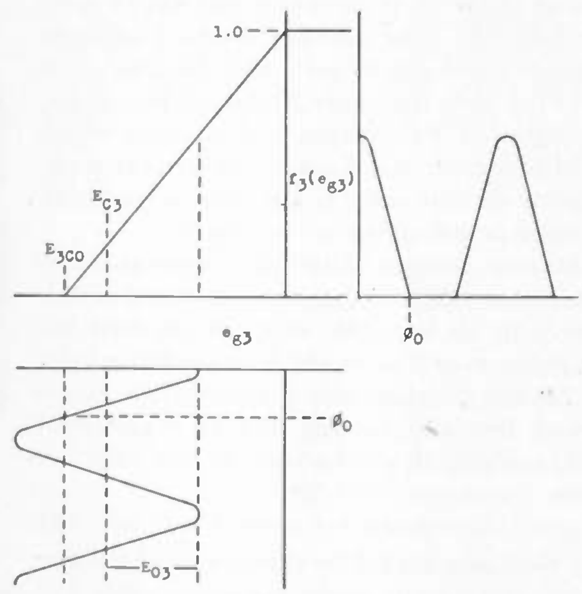

Fig. 15—Wave-form relationship between $f_3(e_{g3})$ and e_{g3} under conditions of Case II.

Operation extends over the region of linear variation into the region of zero plate current. In the first equation below, the subscripts and the superscripts at the end of the brackets indicate that the terms within the brackets are to be evaluated only over that portion of the oscillator cycle lying between the angles indicated. Then

$$f_3(e_{g3}) = \left[\frac{1}{|E_{3co}|}(E_{C3}+E_{O3}\cos\omega_3 t - E_{3CO})\right]_0^{\phi_0} + \Big[0\Big]_{\phi_0}^{\pi}.$$

The first Fourier coefficient is given by

$$a = \frac{1}{|E_{3co}|}\left\{\frac{1}{\pi}(E_{C3} - E_{3CO})\int_0^{\phi_0} d\phi + \frac{1}{\pi}\int_0^{\phi_0} E_{O3}\cos\phi d\phi\right\}$$

and the second coefficient by

$$b = \frac{1}{|E_{3CO}|}\left\{\frac{2}{\pi}(E_{C3} - E_{3CO})\int_0^{\phi_0}\cos\phi d\phi + \frac{2}{\pi}\int_0^{\phi_0}\cos\phi\cos\phi d\phi\right\}.$$

From the above,

$$a = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\sin\phi_0 - \phi_0\cos\phi_0}{\pi}\right\}$$

$$b = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\phi_0 - \sin\phi_0\cos\phi_0}{\pi}\right\}.$$

Case III. Saturation, No Cutoff

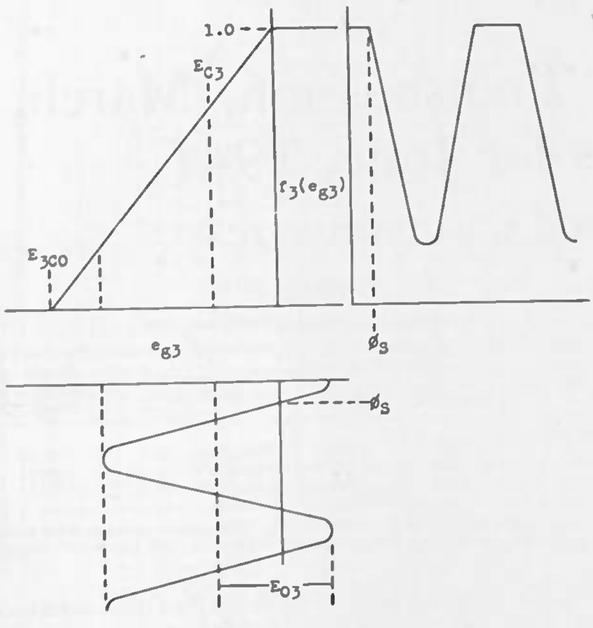

Fig. 16—Wave-form relationship between $f_3(e_{g3})$ and e_{g3} under conditions of Case III.

Operation extends over the region of linear variation into the region of plate-current saturation, where $f_3(e_{g3})$ is constant at the value unity. The equation for $f_3(e_{g3})$ with the terms properly limited is

$$f_3(e_{g3}) = \Big[1\Big]_0^{\phi_S} + \left[\frac{1}{|E_{3OC}|}(E_{C3} + E_{O3}\cos\phi - E_{3CO})\right]_{\phi_S}^{\pi}.$$

From which

$$a = 1 - E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\cos\phi_S + \frac{\sin\phi_S - \phi_S\cos\phi_S}{\pi}\right\}$$

$$b = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{1 - \frac{\phi_S - \sin\phi_S\cos\phi_S}{\pi}\right\}.$$

Case IV. Saturation and Cutoff

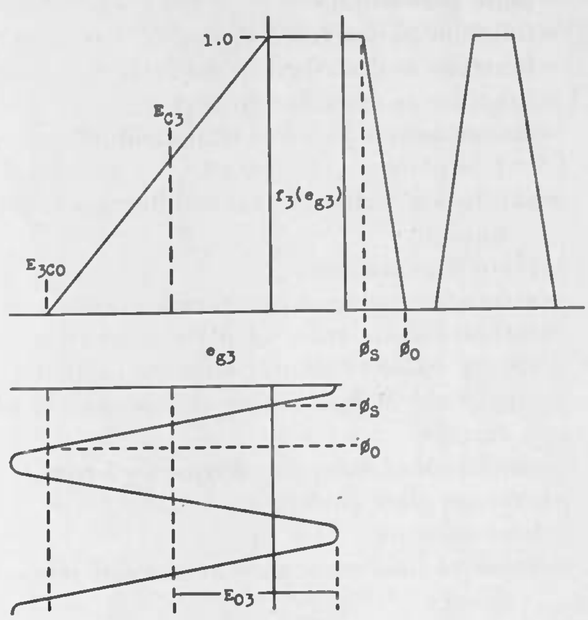

Fig. 17—Wave-form relationship between $f_3(e_{g3})$ and e_{g3} under conditions of Case IV.

Operation extends over the entire region of linear variation of $f_3(e_{g3})$ and into the regions on both sides. For all positive grid-3 voltages, $f_3(e_{g3})$ is constant at the value unity, and for all grid-3 voltages more negative than cutoff voltage, $f_3(e_{g3})$ is constant at the value zero. The equation for $f_3(e_{g3})$ with terms restricted to the proper region is

$$f_3(e_{g3}) = \Big[1\Big]_0^{\phi_S} + \left[\frac{1}{|E_{3CO}|}(E_{C3} + E_{O3}\cos\phi - E_{3CO})\right]_{\phi_S}^{\phi_0} + \Big[0\Big]_{\phi_0}^{\pi};$$

from which

$$a = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\sin\phi_0 - \phi_0\cos\phi_0}{\pi} - \frac{\sin\phi_S - \phi_S\cos\phi_S}{\pi}\right\}$$

$$b = E_{O3}\frac{\mu_{34}}{E_{B4}}\left\{\frac{\phi_0 - \sin\phi_0\cos\phi_0}{\pi} - \frac{\phi_S - \sin\phi_S\cos\phi_S}{\pi}\right\}.$$

APPENDIX II

List of Symbols

a = average value of $f_3(e_{g3})$ as a function of $\omega_3 t$

b = fundamental component of $f_3(e_{g3})$ as a function of $\omega_3 t$

E_{B4} = screen (grid 3 and grid 4) bias voltage

E_{C1} = grid-1 bias voltage

E_{C3} = grid-3 bias voltage

E_{3CO} = grid-3 bias to cause plate-current cutoff

E_{O3} = amplitude of oscillator voltage applied to grid 3

E_{S1} = amplitude of signal voltage applied to grid 1

e_{g1} = instantaneous value of grid-1 voltage

e_{g3} = instantaneous value of grid-3 voltage

e_p = instantaneous value of plate voltage

E_P = plate bias voltage

$f_3(e_{g1})$ = function as described in text

$f_1(e_{g3})$ = function as described in text

$f_p(e_p)$ = function as described in text

g_{1P} = instantaneous value of transconductance, grid 1 to plate

G_{1P} = saturation value of transconductance, grid 1 to plate

g_P = plate conductance

G_{PS} = saturation value of plate conductance

i_p = instantaneous value of plate current

I_{PS} = steady value of saturation plate current

I_P = amplitude of alternating component of plate current

I_d = amplitude of difference-frequency current

I_{Pa} = average plate current

R_L = load resistance

R_d = value of load resistance at the difference frequency

r_P = plate resistance

R_{PS} = saturation value of plate resistance

R_P' = effective plate resistance in converter operation

S_{1P} = conversion conductance, grid 1 to plate

t = time in seconds

ω_1 = 2π times frequency of signal voltage applied to grid 1

ω_3 = 2π times frequency of oscillator voltage applied to grid 3

μ_{34} = cutoff amplification factor of grid 3 with respect to grid 4

Z_L = load impedance

Z_d = value of load impedance at the difference frequency

ϕ = angle of oscillator cycle = $\omega_3 t$

ϕ_0 = angle of oscillator cycle at which cutoff occurs

ϕ_S = angle of oscillator cycle at which saturation conditions cease

The Ionosphere and Radio Transmission, March, 1941, with Predictions for June, 1941*

NATIONAL BUREAU OF STANDARDS, WASHINGTON, D. C.

AVERAGE critical frequencies and virtual heights of the ionospheric layers as observed at Washington, D. C. during March are given in Fig. 1. Critical frequencies for each day of the month are given in Fig. 2. Fig. 3 gives the March average values of maximum usable frequencies, for radio transmission by way of the regular layers. The maximum usable frequencies were determined by the F layer at night

TABLE I

IONOSPHERIC STORMS (APPROXIMATELY IN ORDER OF SEVERITY)

Day and hour E.S.T.	h_F before sunrise (km)	Minimum f_F^0 before sunrise (Mc)	Noon $f_{F_2}^0$ (Mc)	Magnetic character[1] 00–12 G.M.T.	Magnetic character[1] 12–24 G.M.T.	Ionospheric character[2]
March						
30 (from 1300)	—	—	—	1.1	1.7	1.9
31	—[3]	<1.7	<3.9	1.6	0.8	1.8
1	400	<1.7	—[3]	2.0	2.0	1.9
2	—[3]	<1.7	8.9	1.1	0.9	1.3
3	340	diffuse	6.8	0.7	1.0	1.0
4	370	diffuse	8.0	1.0	0.9	0.9
5	324	2.3	7.3	0.7	0.9	0.4
6	345	2.0	6.9	0.7	0.5	0.6
7 (through 0600)	336	1.9	—	0.4	0.5	0.3
13 (from 2100)	—	—	—	0.4	0.6	0.4
14	—[3]	<1.7	<3.9	1.4	1.0	1.7
15	—[3]	<1.7	5.1	0.6	0.5	1.4
16 (through 0600)	317	<1.7	—	0.0	0.0	0.3
28 (from 0500)	350	2.1	7.0	1.1	1.4	0.6
29	298	1.6	7.2	0.9	1.1	0.6
30 (through 1300)	350	<1.7	6.9	1.1	1.7	0.8
19 (from 0700)	—	—	5.9	0.2	0.9	1.2
20 (through 0600)	320	2.1	—	0.8	0.9	0.3
21 (from 2200)	—	—	—	0.8	0.8	0.4
22 (through 0600)	—[3]	1.7	—	1.0	0.7	0.4
For comparison: Average for undisturbed days	301	2.51	8.07	—	—	0.0

[1] American magnetic character figure, based on observations of seven observatories.
[2] An estimate of the severity of the ionospheric storm at Washington on an arbitrary scale of 0 to 2, the character 2 representing the most severe disturbance.
[3] No regular reflections above 2.5 megacycles.

* Report prepared by N. Smith, T. R. Gilliland, A. S. Taylor, F. R. Gracely.

TABLE II

SUDDEN IONOSPHERIC DISTURBANCES

Day	G.M.T. Beginning	G.M.T. End	Locations of transmitters	Relative intensity at minimum[1]
March				
7	1630	1810	Ohio, Ont., D. C.	0.01
20	1650	1810	Ohio, D.C.	0.02
21	1912	2000	Ohio, Ont., D. C.	0.1

[1] Ratio of received field intensity during fade-out to average field intensity before and after, for station W8XAL, 6080 kilocycles, 650 kilometers distant.

TABLE III

APPROXIMATE UPPER LIMIT OF FREQUENCY IN MEGACYCLES OF THE STRONGER SPORADIC-E REFLECTIONS AT VERTICAL INCIDENCE

Day	00	01	02	03	04	05	06	07	08	09	10	11	12	13	14	15	16	17	18	19	20	21	22	23
March																								
1																4	3		3				3	3
2	3			2	3																			
6																						5		3
7	3								3															
9																						3		
11																								3
12			3																					
21																				3				
22					3													3						
24																		3	3					
26								3																
28																				3	3		3	
31		5																						

and by the F_2 layer during the day. Fig. 4 gives the expected values of the maximum usable frequencies for radio transmission by way of the regular layers, average for undisturbed days, for June, 1941. All of the foregoing are based on the Washington ionospheric observations, checked by quantitative observations of long-distance reception.

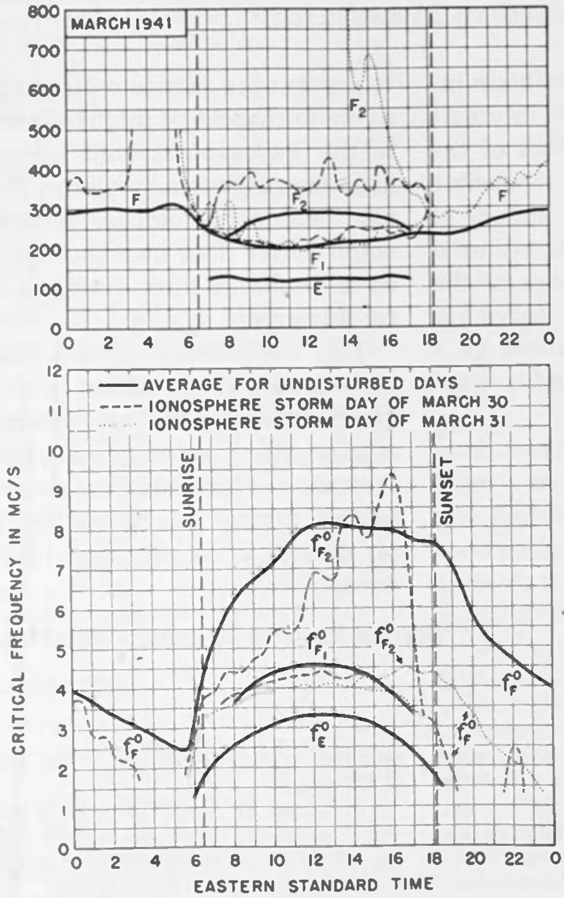

Fig. 1—Virtual heights and critical frequencies of the ionospheric layers, observed at Washington, D. C., March, 1941. The gaps in the dashed and dotted graphs indicate periods on the ionospheric storm days of March 30 and 31 when the $f_F{}^0$ was below the limit of the recorder.

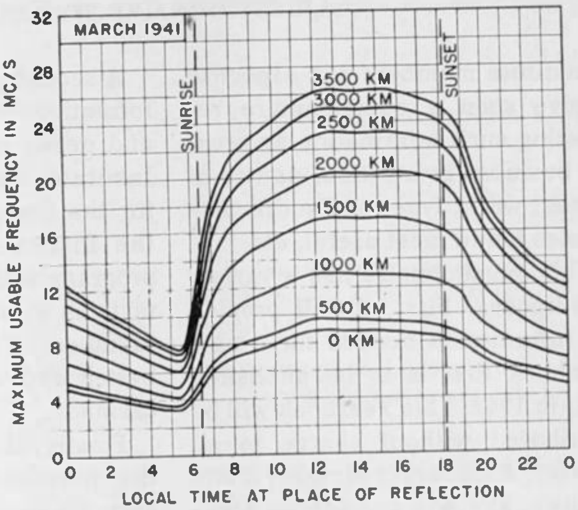

Fig. 3—Maximum usable frequencies for dependable radio transmission via the regular layers, average for March, 1941. These curves and those of Fig. 4 also give skip distances, since the maximum usable frequency for a given distance is the frequency for which that distance is the skip distance.

Ionospheric storms are listed in Table I. The details of the ionospheric storm days of March 30 and 31 are shown in Fig. 1. The open circles in Fig. 2 indicate the noon and midnight critical frequencies observed during the ionospheric storms listed in Table I. The sizes of the circles roughly represent the severity of the storm.

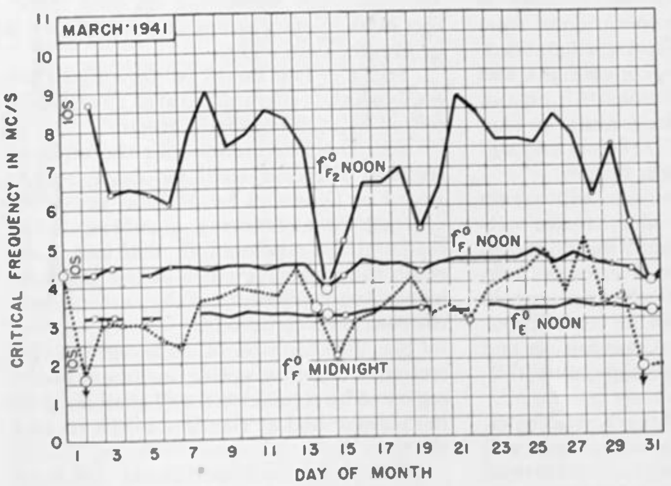

Fig. 2—Midnight $f_F{}^0$ and noon $f_E{}^0$, $f_{F_1}{}^0$, and $f_{F_2}{}^0$ for each day of March. Open circles indicate critical frequencies observed during ionospheric storms; sizes of circles represent approximately the severity of the storms. The letters "IOS" on March 1 indicate no reflections observed during a severe ionosphere storm. Arrows indicate midnight critical frequencies of less than 1.7 megacycles observed during the nights of March 1–2 and 30–31.

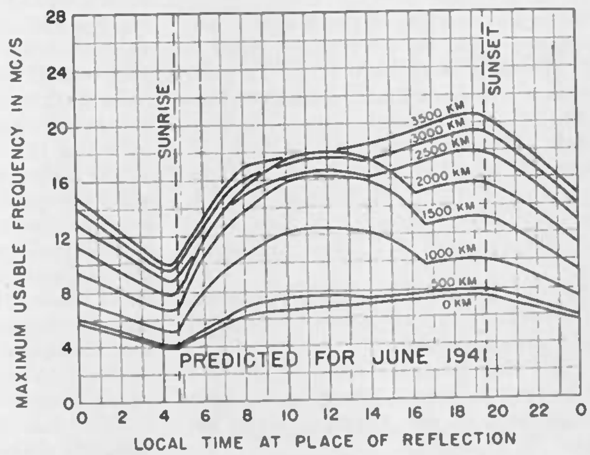

Fig. 4—Predicted maximum usable frequencies for dependable radio transmission via the regular layers, average for undisturbed days, for June, 1941. The values shown will be considerably exceeded during irregular periods by reflections from clouds of sporadic E layer. For information on use in practical radio transmission problems, see Letter Circulars 614 and 615 obtainable from The National Bureau of Standards, Washington, D. C., on request.

The sudden ionospheric disturbances are listed in Table II. Table III gives the approximate upper limit of frequency of strong sporadic-E reflections at vertical incidence.

Institute News and Radio Notes

Questionnaire to Serve the Institute Membership

Institute members can expect to receive soon a questionnaire requesting such information as present business connection, nature of present work, types of Proceedings articles found most useful, etc.

This questionnaire has a variety of purposes. First, it will provide the information needed for a new Yearbook that is to be published early in 1942. This Yearbook will be distributed without charge to all Juniors, Associates, Members, and Fellows, and will provide an accurate up-to-date list of the members of the Institute, with their commercial connections and business titles.

A second purpose is to obtain information on the technical interests and needs of the members of the Institute. This will make it possible for the first time in the history of the Institute to plan an editorial program with a knowledge of the facts as to what will best serve our members. The result should be a better and a more valuable Proceedings.

Finally, the facts obtained from the questionnaire will give a more accurate picture of what the Institute really is than has ever before been obtained. This will make it possible to present the Institute more effectively to those who might advertise in the Proceedings, and hence to increase advertising revenue. It will also enable the officers to understand better their own society, and so to discharge their duties to better advantage.

The co-operation of the Institute members in filling out the questionnaire *in full*, and returning it promptly is urgently solicited. A few minutes time and thought given to this matter will ultimately pay big dividends on your membership dues.

Frederick Emmons Terman, President

Board of Directors

A regular meeting of the Board of Directors was held on Wednesday, April 2, 1941. Those present were F. E. Terman, president; Haraden Pratt, treasurer; Austin Bailey, A. B. Chamberlain, I. S. Coggeshall, Melville Eastham, Alfred N. Goldsmith, Virgil M. Graham, R. A. Heising, L. C. F. Horle, C. M. Jansky, Jr., B. J. Thompson, H. M. Turner, A. F. Van Dyck, H. A. Wheeler, and H. P. Westman, secretary.

Seventy-six applications for Associate, six for Junior, and forty-seven for Student grades were approved.

The Pacific Coast Convention for 1941 will be held in Seattle, Washington, with headquarters at the Olympic Hotel on August 20, 21, and 22.

The Committee on Special Papers reported that it was making substantial progress in the obtaining of broad survey or tutorial types of papers for the Proceedings.

The secretary was instructed to omit the names of Students from the Yearbook which is to be published early in 1942.

Since its establishment a little over a year ago, the Executive Committee has devoted its time primarily to a preliminary consideration of the same matters which came before the entire Board of Directors at the regular monthly meetings. The Executive Committee held no authority and it could not, therefore, do very much in relieving the Board of Directors of the handling of routine matters. The experience gained under this form of operation now permits the establishment of a specified scope of activities between the Executive Committee and the Board of Directors. Such a distribution of powers and duties were approved by the Board of Directors.

Under the new arrangement, the Executive Committee will be comprised of the president, treasurer, secretary, editor, and two or three Board members. The individual members of the Executive Committee will be assigned certain specific fields for which they will be responsible.

Broadly, the Executive Committee will be responsible for the routine management of the Institute. It will have charge of the appointment and operation of most of the Institute committees, all publications, the Institute office, conventions, and the establishment of sections. All acts of the Executive Committee must be approved by the Board of Directors and will be made acts of the Board.

The Board of Directors will retain complete authority over all Institute matters. It will determine and formulate all policies under which the Executive Committee will operate. It will continue in direct charge of the following committees: Appointments, Awards, Constitution and Laws, Executive, Nominations, and Tellers.

As a result of these changes, it is anticipated that the Institute's management and leadership will be substantially improved and the Board of Directors will have the time to consider in greater detail the fundamental problems which face the Institute.

In order to permit more widespread participation in the management of the Institute, the establishment of regional Directors is contemplated. It is expected that it will be necessary for the Institute to defray in part the expenses of these regional Directors incidental to their attendance at certain meetings of the Board of Directors.

Provision for such Directors must be voted into the Constitution and the following proposed amendments were adopted for submission to the membership.

Sec. A—In addition to the Officers of the Institute provided for in Article V, Sec. 1 of the Constitution there is hereby established the office of regional Director.

Sec. B—The Board of Directors shall establish by Bylaw the number of regional Directors. The number of regional Directors holding office at any one time shall not be greater than eight.

Sec. C—The Board of Directors shall establish the geographical area to be represented by each regional Director.

Sec. D—Nominations for the office of regional Director shall be made to the Board of Directors by petition signed by at least twenty voting members of the region to be represented. Petitions must reach the executive office of the Institute before August fifteenth of the year in which the election is to be held.

Sec. E—Election of each regional Director shall be by voting members in the region to be represented and shall occur in the manner and at the time of the regular election.

Sec. F—Candidates for the office of regional Director must be Members or Fellows of the Institute and must not have been members of the Board of Directors within one year of the beginning of the term of office to be filled.

Sec. G—Each regional Director shall be a member *ex officio* of the Board of Directors of the Institute.

Sec. H—The term of office of each regional Director shall be two years.

Committees

Admissions

A meeting of the Admissions Committee was held on December 4. Those present were A. F. Van Dyck, chairman; Ralph Bown, F. W. Cunningham, C. W. Horn, C. M. Jansky, Jr., H. B. Richmond, and H. P. Westman, secretary. Thirteen applications for transfer to Member grade were considered. Of these, eleven were approved, one was denied, and one was tabled. Of six applications for admission to Member grade, five were approved, and one was denied.

Board of Editors Co-ordinating Committee

The Co-ordinating Committee of the Board of Editors met on December 27. Alfred N. Goldsmith, chairman; P. S. Carter, H. A. Wheeler, Helen M. Stote, assistant editor; and H. P. Westman, secretary, were present. The committee reviewed a number of papers which were examined previously by the Papers Committee and Board of Editors and rendered decisions on their acceptability for publication in the PROCEEDINGS.

Convention

Meetings of the Convention Committee were held on November 19 and on December 12. At the earlier meeting, the attendance consisted of H. P. Westman, chairman and secretary; A. B. Chamberlain, I. S. Coggeshall, E. J. Content (representing J. R. Poppele), J. D. Crawford, assistant secretary, D. D. Israel, C. E. Scholz, Helen M. Stote, Lincoln Walsh, and William Wilson.

The December meeting was attended by H. P. Westman, chairman and secretary; I. S. Coggeshall, E. K. Cohan, E. J. Content (representing J. R. Poppele), J. D. Crawford, assistant secretary; D. D. Israel, C. E. Scholz, and Helen M. Stote. These meetings were devoted to the preparation of the program arrangements for the annual convention which was held in New York City on January 9, 10, and 11.

Membership

E. D. Cook, chairman; I. S. Coggeshall, E. J. Content, F. W. Cunningham, L. G. Pacent, C. R. Rowe, Bernard Salzberg, and J. D. Crawford, secretary to the committee; attended a meeting of the Membership Committee which was held on December 4. Preliminary arrangements were made for the preparation of a list of Associates whose transfer to Member grade would be recommended to the Admissions Committee and the Board of Directors.

New York Program and Sections

A proposal was made to the Board of Directors that a New York Section of the Institute be established. The Board of Directors referred this matter jointly to the New York Program Committee and to those members of the Sections Committee appointed directly by the President. Those who were present at the meeting on December 27 were H. P. Westman, acting chairman and secretary; J. H. Miller, chairman of the Sections Committee; Austin Bailey, N. P. Case, I. S. Coggeshall, G. C. Connor, E. D. Cook, W. M. Goodall, Wallace James, G. T. Royden, C. E. Scholz, and J. D. Crawford, assistant secretary.

A report which was unfavorable toward the establishment of a New York Section was prepared. In addition, arrangements for several future meetings in New York were discussed.

COMING MEETINGS

Summer Convention

Detroit, Michigan

June 23, 24, and 25, 1941

Emporium Section

Summer Seminar

August 1 and 2, 1941

Public Relations

The Public Relations Committee met in Rochester, New York, on November 11. Those present were Virgil M. Graham, chairman; L. C. F. Horle, president; Melville Eastham, treasurer; W. R. G. Baker (guest), J. E. Brown, E. T. Dickey (representing E. W. Engstrom), A. L. Durkee (representing G. W. Gilman), D. G. Fink, George Grammer, G. E. Gustafson (guest), R. A. Hackbusch, D. D. Israel, I. J. Kaar, R. H. Manson, A. F. Van Dyck, Craig Walsh (representing Lincoln Walsh), H. A. Wheeler, R. J. Wise (guest), and H. P. Westman, secretary.

On November 27, a meeting of the committee was held at which Virgil M. Graham, chairman; E. T. Dickey (representing E. W. Engstrom), A. L. Durkee (representing G. W. Gilman), D. G. Fink, D. D. Israel, A. F. Van Dyck, Lincoln Walsh, J. D. Crawford, assistant secretary, and H. P. Westman, secretary, were present.

In December, on the 18th, a meeting of this committee was attended by Virgil M. Graham, chairman; G. W. Gilman, D. D. Israel, A. F. Van Dyck, Lincoln Walsh, H. A. Wheeler, J. D. Crawford, assistant secretary; and H. P. Westman, secretary.

The major matters considered by this committee concerned the program for the annual banquet held on January 10, during the annual convention in New York, the preparation of news releases on current items of interest, and the preparation of a special release in advance of the publication in the PROCEEDINGS of the annual review of developments during 1940.

Technical Committees

Annual Review

A meeting on November 22 and another on December 16 were held to develop methods for the preparation of the annual review for 1940. Both of these meetings were attended by L. E. Whittemore, chairman of the Annual Review Committee; Keith Henney, H. A. Wheeler, chairman of the Standards Committee; J. D. Crawford, secretary to the technical committees; and H. P. Westman, secretary of the Institute.

Electronics

The Technical Committee on Electronics met in Rochester, N. Y., on November 12 and those present were P. T. Weeks, chairman; H. A. Wheeler, chairman of the Standards Committee; R. L. Freeman, E. C. Homer (representing H. P. Corwith), S. B. Ingram (representing J. R. Wilson), Ben Kievit, Jr., D. W. Jenks (representing K. C. DeWalt), G. F. Metcalf, G. D. O'Neill, H. W. Parker, and J. D. Crawford, secretary to the committee.

The committee examined into the progress being made by the various subcommittees in the preparation of annual-review material and also in regard to standardization activities.

Subcommittee on High-frequency Tubes

F. B. Llewellyn, chairman; R. L. Freeman, L. S. Nergaard, and J. D. Crawford, secretary to the committee; attended a meeting of the Subcommittee on High-Frequency Tubes of the Electronics Committee, which was held on November 28. The committee devoted its efforts to a consideration of a number of definitions which had been referred to it by the Electronics Committee.

Subcommittee on Photoelectric Devices

The Subcommittee on Photoelectric Devices of the Technical Committee on Electronics met in Rochester, N. Y., on November 11. Those present were Ben Kievit, Jr., chairman; L. W. Morrison (representing E. F. Kingsbury), A. D. Power (representing A. M. Glover), and H. P. Westman, secretary. Matters concerning both standardization and the preparation of annual-review material were considered.

Subcommittee on Small High-Vacuum Tubes

This subcommittee of the Technical Committee on Electronics met on November 26 and those present were R. S. Burnap, chairman; G. W. Bain, E. H. Erickson (guest), R. H. Fidler, R. C. Hergenrother, G. D. O'Neill, E. A. Veazie, and J. D. Crawford, secretary to the committee. Its report for the Annual Review Committee was prepared and it considered also a number of items pertaining to both definitions and testing of small high-vacuum tubes.

Electronics Conference

The committee in charge of the Electronics Conference met on November 7 and those present were F. R. Lack, chairman; F. B. Llewellyn, A. L. Samuel, R. W.

Sears, B. J. Thompson, and J. D. Crawford, secretary to the committee. This meeting was devoted to an analysis of the conference which had been held during the previous month. A number of recommendations which may be useful to the committee preparing for the next conference were made.

Facsimile

The Technical Committee on Facsimile met on November 25. Those present were J. L. Callahan, chairman; W. A. R. Brown, E. S. Fergusson (representing Robert Hatch), J. H. Hackenberg (representing J. W. Milnor), C. W. Harrison (representing J. R. Poppele), J. V. L. Hogan, H. C. Knutson, R. E. Mathes, P. Mertz, Frank Turner (representing John Hancock), H. J. Tyzzer (representing W. G. H. Finch), C. J. Young, and J. D. Crawford, secretary to the committee. The time was devoted to a consideration of definitions in the facsimile field.

Subcommittee on Definition Grouping

To avoid using much time of the main committee in making an orderly arrangement of the definitions on facsimile, a special subcommittee was appointed. That committee met on November 29 and those present were H. C. Knutson, chairman; E. S. Fergusson (representing Robert Hatch), R. E. Mathes, H. J. Tyzzer (representing W. G. H. Finch), and J. D. Crawford, secretary to the committee.

Frequency Modulation

The Technical Committee on Frequency Modulation met in Rochester N. Y., on November 11. Those present were H. A. Wheeler, chairman of the Standards Committee and acting chairman of the meeting; S. L. Bailey (representing C. M. Jansky, Jr.), R. C. Ballard (guest), G. T. Bennett (guest), J. E. Brown, H. B. Canon (guest), K. A. Chittick (guest), E. D. Cook (representing H. B. Marvin), M. G. Crosby, R. D. Duncan (representing C. B. Jolliffe), A. L. Durkee (representing G. W. Gilman), P. F. G. Holst (guest), K. W. Jarvis (guest), J. K. Johnson (guest), J. D. Parker (representing A. B. Chamberlain), L. P. Wheeler, and J. D. Crawford, secretary to the committee. The main activities of the meeting were in the consideration of definitions. A continuation of this work occurred at a meeting on December 9 at which were present D. E. Noble, chairman; J. G. Chaffee (representing G. W. Gilman), W. H. Moffat (representing A. B. Chamberlain), H. A. Wheeler, chairman of the Standards Committee; and J. D. Crawford, secretary to the committee.

Subcommittee on Definitions

Two meetings of the Subcommittee on Definitions of the Technical Committee on Frequency Modulation were held during December. At the December 7 meeting, C. C. Chambers, chairman; M. G. Crosby, G. W. Gilman, H. A. Wheeler, chairman of the Standards Committee; and J. D. Crawford, secretary to the committee; were present. The meeting on the 23rd was attended by C. C. Chambers, chairman; J. G. Chaffee (representing G. W. Gilman), M. G. Crosby, H. A. Wheeler, chairman of the Standards Committee; and J. D. Crawford secretary to the committee.

Radio Receivers

Subcommittee on Frequency-Modulated-Wave Receivers

R. M. Wilmottee, chairman; D. E. Foster, chairman of the Technical Committee on Radio Receivers; A. W. Barber, R. I. Cole, L. F. Curtis, M. L. Levy (representing W. F. Cotter), J. A. Worcester (representing W. M. Angus), and J. D. Crawford, secretary to the committee; attended a meeting on December 2 of the Subcommittee on Frequency-Modulated-Wave Receivers of the Technical Committee on Radio Receivers. The meeting was devoted to proposals for the testing of these receivers and the equipment used in such tests.

Television

A meeting on November 11 in Rochester, N. Y., of the Technical Committee on Television was attended by I. J. Kaar, chairman; H. A. Wheeler, chairman of the Standards Committee; R. C. Ballard (guest), J. E. Brown, T. J. Buzalski (guest), K. A. Chittick (representing E. W. Engstrom), D. E. Foster, D. E. Harnett (representing A. V. Loughren), A. G. Jensen, R. H. Manson (guest), Harry Sadenwater (guest), R. E. Shelby, A. E. Thiessen (representing D. B. Sinclair), W. B. Whalley (guest), and J. D. Crawford, secretary to the committee. Work on both annual-review and standards activities was consummated.

Subcommittee on Transmission Lines and Antennas

L. M. Leeds, chairman; C. R. Burrows, R. F. Lewis, N. E. Lindenblad, D. B. Sinclair, and J. D. Crawford, secretary to the committee; attended a meeting of the Subcommittee on Transmission Lines and Antennas of the Technical Committee on Television, which was held on December 13. This committee is preparing material on the subjects indicated by its title for inclusion in the television standards report.

Sections

Atlanta

A paper on "Audio-Frequency Measurements" was presented by Walter Van Nostrand, of the Van Nostrand Engineering Service.

In measuring the frequency of a radio transmission, use is made of a heterodyne frequency meter and a crystal-controlled oscillator driving multivibrators. Methods of interpolation between points of the heterodyne frequency meter through the use of the harmonic output of the multivibrators were discussed. The use of an audio-frequency oscillator as the final step in matching frequencies was described.

In measuring signals which are available for short periods of time only, the audio-frequency oscillator is adjusted to zero beat with the heterodyne note between the standard signal and the signal to be measured. This permits the frequency of the audio-frequency oscillator to be measured when the signal has stopped.

A resistance-tuned audio-frequency oscillator and an electronic-type audio-frequency meter which are used in these measurements were described. The frequency meter is of the direct-reading type. In the discussion of the paper, Major Van Nostrand gave additional information on the construction of the audio-frequency oscillator and meter.

This was the annual meeting of the section and A. W. Shropshire, WSB, was elected chairman; J. M. Comer, Jr., WATL, Atlanta Broadcasting Company, was named vice chairman; and G. M. Howard, was designated secretary-treasurer.

January 17, 1941, P. C. Bangs, chairman, presiding.

Baltimore

The paper on "The Measurement of Coil Reactance in the 100-Megacycle Region" by F. Hamburger and C. F. Miller, which was published in the October issue of the PROCEEDINGS, was presented at this meeting.

March 21, 1941, J. E. Allen, chairman of the Papers Committee, presiding.

Buffalo-Niagara

P. S. Christaldi, of the Research Department of the Allen B. DuMont Laboratories, presented a paper on "The Use of Cathode-Ray Tubes in Commercial Measurements."

A description of various types of cathode-ray tubes was first presented. The control of the size and intensity of the spot through the design of the tube and by means of varying the applied voltages were discussed. Deflection sensitivity was next considered and the effect of the construction of the tube on this characteristic was outlined.

Four general types of screens are used. The most popular for oscillographic work gives a green light. For photography, a blue characteristic is preferred. If transients are to be studied, a long-persistence screen which is usually of green color is required. Television utilizes a medium-persistence characteristic and white light.

It was pointed out that the deflection of the spot indicates the peak voltage applied to the deflecting plates. Some tubes are made with special electrodes at the screen end of the tube to permit the synchronizing of mechanical movements and for use as relays. Newer tubes and circuits have increased the usefulness of these devices in many fields.

March 12, 1941, B. Atwood, chairman, presiding.

Cincinnati

R. J. Rockwell, chief engineer of WLW, WSAI, and WLWO, presented a paper on

"Modernizing a Broadcast Transmitter."

Most broadcast stations either purchase new equipment about every five years or make continuous changes in their equipment as improved methods or apparatus become available.

If the equipment is replaced every five years, a modern appearance is easily retained, no engineering staff will be required as this service will be obtained from the supplier of the equipment, and a lower cost will result in small stations because the equipment cost is not large.

Among the disadvantages of this system will be found the necessity of the operating staff requiring a substantial period of time in which to become familiar with the new equipment, operating costs during the life of the equipment may be higher than necessary because improved equipment and methods are not used, difficulties may be encountered in the changeover when the space occupied by the old equipment may be required for the new, and the old equipment is sually disposed of at a financial loss.

When continuous improvements are made, new developments may be applied with a minimum loss of time, maintenance of the equipment may be handled more effectively, the equipment may be adapted to the particular need and physical surroundings, and, where several transmitters are being used, common transmission facilities may be easily adapted to their joint requirements.

The equipment under the control of the author is improved continuously. A description was given of the mechanical and electrical changes made in the 5-kilowatt WSAI transmitter after it was removed from the center of Cincinnati to Mt. Healthy, Ohio, which is approximately 12 miles north of the downtown section of the city.

Particular stress was given to the improvements in negative feedback which were obtained by eliminating those elements in the feedback circuit which produce phase rotation. In the mechanical redesign of the transmitter, emphasis was placed on the safety features. All cabinets containing high voltages have grounding systems on each door which operate immediately on opening. The old installation used a complicated arrangement which resulted in a substantial loss in time.

Pictures were shown of each redesigned transmitter unit and of the ventilating system used in the new building which houses these units. Motion pictures were shown of the construction of the new building and erection of the three half-wave tower antennas.

The meeting was concluded by the showing of motion pictures of the construction of an amateur rotating beam antenna which was made by T. P. Jordan.

March 18, 1941, J. M. McDonald, chairman, presiding.

Cleveland

"Interference Problems in Fluorescent Lighting" was the subject of a paper by E. S. Mills and J. H. Campbell of Nela Park.

Mr. Mills first described fluorescent lamps and their light characteristics. The various colors which are available in fluorescent lamps were then demonstrated. It was stated that the efficiency increases with the length of the lamp and that for a given input power, it is about three times that of an incandescent lamp.

Mr. Campbell then described the circuits used with them. In addition to manual, automatic and thermal switches are employed to start the lamp. Autotransformers to step up the line voltage are required for some lamps. Ballast resistors are used in series with them. He demonstrated by means of a rotating disk the stroboscopic effect of a single lamp and that the use of two lamps operated out of phase with each other are practically as good as an incandescent lamp.

Noise interference reaches radio receivers by three principal channels. They are direct radiation from the bulb to the radio receiver, radiation from the power line to the radio receiver, and propagation over a power line which is common to both devices. The noise ranges in frequency from 150 to 1800 kilocycles, peaking around 800 to 900 kilocycles. Simple capacitive filters are usually effective. Suppressors and additional filters may be required if the noise must be reduced to a very low value. The disturbance appears to originate at the cathode rather than in the positive column of the lamp.

A portable receiver was used to demonstrate the noise from a lamp, the intensity of the noise being indicated on a meter. It was stated that the wattage of the lamp has little to do with the noise generated.

February 27, 1941, C. E. Smith, chairman, presiding.

Connecticut Valley

"The Connecticut State Police System of Frequency-Modulation Communication" was the subject of a paper by S. E. Warner, supervisor of maintenance for the Connecticut State Police Radio.

Before development work on the system was started, a choice had to be made between a low-frequency amplitude-modulated-wave system providing one-way communication and a high-frequency frequency-modulated-wave system providing two-way communication. Professor Noble of Connecticut State College recommended the latter system and in a period of one year the development work and complete installation was finished.

All automobile receivers operate on the same frequency, the more powerful signal being in control. Interference from transmitters located in the various barracks is reduced by requiring them to wait their turn before going on the air.

Many of the cars are equipped to use the main-station frequency for emergency service. If all the main stations should become inoperative simultaneously, some of these cars would be directed to operate from hilltops using the main-station frequency until the regular stations resumed operation.

The cars transmit normally on a different frequency than the main station. This prevents their transmissions from being blanketed by the operation of a main station in an adjacent territory.

Pre-emphasis is employed and a gain of 12 decibels is obtained at 3000 cycles. This provides reasonably constant amplitude for male voices.

Distortion is governed somewhat by the multiplication of the original oscillator frequency. This is 32 times and is a compromise between cost and distortion.

A squelch circuit operates on less than 1 microvolt and uses 2 limiters. General noise reduction has been found to be less important than the reduction of impulse noise which for values of 50 to 100 times the signal can be reduced to the signal level by the limiter. The greatest advantage of frequency modulation over amplitude modulation is evident when the signal-to-noise ratio is very bad. Under these conditions, the frequency-modulated signal is fully intelligible when the amplitude-modulated signal is a total loss.

In the discussion, the antenna design was treated. To determine the effect of standing waves on the operation of the coaxial feed line, the region inside the line was explored with an insulated probe. It was found that the distance above ground or the antenna height had little effect. The skirt dimensions were critical and proper design resulted in an improvement in operation.

February 20, 1941, K. A. McLeod, chairman, presiding.

"Beam Power Tetrodes as Negative-Resistance Oscillators" was the subject of a paper by H. L. Krauss, graduate student at Yale University.

The relations between plate and screen current in a beam power tetrode, 6L6, were first described. The circuit employed for oscillation is similar to that of the transitron, except that the tuned circuit is placed between the screen and plate and the voltages applied to these electrodes through the center-tapped inductance of the tuned circuit.

Tube structures were shown and the voltage gradients under various instantaneous conditions were outlined. The amount of negative resistance varies inversely with the plate voltage and directly with the grid bias. The requirements for oscillation were outlined.

By means of a cathode-ray oscillograph, data on the operation of the circuit were obtained. These measurements were simplified by driving the circuit with 60-cycle voltages. The effect of changes in electrode voltages on the purity of the output of the oscillator was discussed. Also, the necessity of obtaining a proper value of the negative resistance in relation to the circuit characteristics was pointed out.

A series of curves were shown of the values of negative resistance at many possible operating conditions. As the negative resistance is developed in the plate circuit, the impedance of the power supply is of considerable importance and must be kept to a very low value by the use of some sort of a regulator.

The uses of this oscillator, methods of modulating it, and the efficiency of operation were then discussed.

March 11, 1941, W. M. Smith, secretary-treasurer, presiding.

Dallas-Fort Worth

"Feedback in Audio-Frequency Amplifiers" was the subject of a paper by C. L. Farrar, professor of electrical engineering at the University of Oklahoma.

The need for reducing distortion and noise in audio-frequency amplifiers and the use of inverse feedback to accomplish this were first discussed.

By means of a cathode-ray oscillograph, the distortion in the output of an audio-frequency amplifier, operated from a single-frequency oscillator, was shown. Inverse feedback could then be applied and varied to demonstrate its effect.

By using various combinations of resistance and capacitance in the feedback circuit, the characteristic of the amplifier could be modified. The demonstration amplifier had an output of 15 watts and with 17 decibels of feedback had less than 1 per cent distortion and a noise level of −70 decibels.

It was pointed out that in a radio transmitter, inverse feedback from the output of the transmitter back to the audio-frequency stages requires that the radio-frequency stages have a 35-kilocycle band width if feedback of 20 decibels is to be employed.

March 21, 1941, T. L. Kimzey, vice chairman, presiding.

Emporium

F. M. Budelman, chief engineer of F. M. Link Company, presented a paper on "Design Problems of Mobile Equipment." It concerned the frequency-modulation communication system used by the Connecticut State Police. It was first pointed out why frequency modulation was chosen in preference to amplitude modulation.

A modification of the Armstrong phase modulation was used in the transmitter. The audio-frequency range covers from 500 to 3000 cycles. This is most effective for the transmission of the male voice when intelligibility is the prime consideration. The mobile units used in the police cars give an output of 25 watts.

The receiver is a high-gain superheterodyne using two limiters to obtain maximum noise reduction. A squelch circuit reduces the noise output of the receiver in the car when the signal is not being received.

Illustrations of the installation were shown and one of the mobile units was available for examination.

In addition to this police application, other uses of these equipments for army, navy, and coast guard services were outlined.

March 13, 1941, R. K. Gessford, chairman, presiding.

Indianapolis

B. V. K. French, radio engineer for P. R. Mallory and Company, presented a paper on "Application of Inductive Tuning to Ultra-High Frequencies."

The system described employed inductive tuning for a frequency range extending from 20 to approximately 200 megacycles. It employs a continuously variable slide-wire inductor. The system provides high reset accuracy, immunity from the effects of vibration, low microphonism, and an extended frequency range. It is especially applicable to the aircraft and military field as well as for instrument purposes.

At this meeting an election of officers took place and A. N. Curtiss of the RCA Manufacturing Company, was elected chairman; S. E. Benson, of Farnsworth Television and Radio Corporation, was named vice chairman; and T. N. Rosser, of P. R. Mallory and Company, was named secretary-treasurer.

March 21, 1941, A. N. Curtiss, chairman, presiding.

Pittsburgh

V. K. Zworykin, director of electronic research, RCA Manufacturing Company, presented a paper entitled "Image Formation by Electrons."

The history of electron optics was traced from the discovery of the focusing effect of space charges to the development of the electron microscope. The close parallel between light and electron optics was pointed out and the construction of lenses was discussed. Electron lenses are subject to aberrations similar to glass lenses and the technique of correcting for some of these effects was described.

The fundamental limitation on the resolving power of the light microscope is the wavelength of the light. Some gain can be obtained by using ultraviolet light. However, this gain is relatively slight compared with the gain in resolution obtained by using high-velocity electrons.

The design of the electron microscope is dependent upon a knowledge of the distribution of the focusing fields and the trajectories of the electrons. The electrolytic tank used in determining field distribution and the method of predicting electron trajectories were described.

The method of mounting specimens was described and slides showing the results obtainable with the electron microscope were shown.

This meeting was held jointly with the Carnegie Institute of Technology Chapter of Sigma Xi.

March 8, 1941, Dr. Fettke, president, Sigma Xi Chapter, presiding.

R. K. Crooks, engineer for the Union Switch and Signal Company, presented a paper on "Amplifier Response to Square Waves."

The principles involved in the testing of amplifiers by means of square waves were discussed. It was pointed out that the method yields essentially qualitative information regarding amplifier response. However, quantitative information can be obtained under certain conditions.

By means of a graphical representation of the Fourier series representing a square-wave function, the effect of changes in the amplitude and phase relations of the various components in the series was shown. Such changes in the amplitude or phase relations of the components in the impressed wave may occur in an amplifier and are the basis of square-wave testing.

The three major types of distortion which may occur in amplifiers and the effect of each upon the observed output wave with a square-wave input were discussed. These points were demonstrated on four typical amplifiers. A square-wave voltage was impressed upon the amplifier input in question and the output voltage observed on an oscilloscope. The effect of phase and frequency distortion was observed in the output wave. The effect of amplitude distortion was not demonstrated but the results of such distortion were discussed.

March 24, 1941, R. E. Stark, chairman, presiding.

Portland

"Antennas for C.A.A. Transoceanic Communication Systems" was the subject of a paper by Sydney Pickles, radio engineer for the Civil Aeronautics Authority.

Transoceanic, meteorological, and aircraft communication services are required of the stations using these antennas. Practically all types of high-frequency radiating systems are required and vary from high-gain highly directive structures to nondirective systems for local broadcasts. Multielement arrays, simple doublets, rhombics, and V antennas are employed.

To conserve space, nearly all of the antennas are used at two or more frequencies, the frequencies differing by about 25 per cent to compensate for the varying conditions met along the transmission path.

The design of arrays to provide considerable gain over such wide frequency ranges was then discussed. The basic element is a simple doublet of 5/4 wavelengths at the highest frequency to be radiated. This gives a gain of approximately 6 decibels over that of a half-wave doublet. A gain of 15 decibels can be obtained by separating two of these antennas by 1.84 wavelengths between centers and with a duplicate set placed 0.65 wavelength directly above the first. All radiators are accompanied by reflectors spaced 0.25 wavelength to the rear. The mid-point between the upper and lower elements was shown to be 1 wavelength above ground.

The directivity at both horizontal and vertical axes of such a structure is satisfactory and a loss in gain of approximately 3 decibels results at the lower transmission frequency. When supported on 90-foot poles, the upper and lower radiating elements ceased to be desirable at frequencies much below 8 megacycles. At the lower frequencies the use of only the higher elements was shown to be advisable.

March 6, 1941, E. R. Meissner, chairman, presiding.

P. C. Sandretto, director of the communications laboratory of United Air Lines, presented a paper on "Radio Aids to Avigation."

A second paper on "Frequency-Modulation Systems and Methods" was pre-

sented by E. S. Winlund, radio engineer for the RCA Manufacturing Company.

March 19, 1941, E. R. Meissner, chairman, presiding.

Rochester

F. S. Goucher of the Bell Telephone Laboratories presented a paper on "The Microphone and Research."

The evolution of the carbon microphone was traced from the earliest experimental models to the present commercial types. By means of a magnetic tape recorder, the quality of reproduction and efficiency of various models were demonstrated.

It was shown that the nature of microphonic action in contacts depends on the elastic deformation of minute hills or roughnesses on the contact surfaces and is affected by exceedingly minute motions between the contact particles. A microphonic contact operated through a section of a steel rail acting as a diaphragm showed clearly how small a motion is required to obtain an effective variation in the resistance of the contact.

The meeting was held jointly with the local section of the American Institute of Electrical Engineers, the Optical Society of America, and the Rochester Engineering Society.

March 6, 1941, O. L. Angevine, Jr., secretary, presiding.

San Francisco

E. S. Winlund, transmitter engineer for the RCA Manufacturing Company, presented a paper on "Frequency Modulation—Transmitting Circuits and Design."

The relative characteristics of the Armstrong, Crosby, RCA, and Western Electric frequency-modulation transmitter circuits were given. A detailed description of the RCA circuit was then presented. A feature of this is that the discriminator and modulated-oscillator tank circuits are enclosed in a dual heat oven. As a result of this and other features, the frequency stability characteristics are very good. Under tests of extreme conditions, frequency variation was only 700 cycles at 42 megacycles; 4000 cycles is permitted.

Two audio-frequency inputs are provided, one following the Radio Manufactures Associatian pre-emphasis curve from 30 to 15,000 cycles, and the other being flat to 25,000 cycles. Distortion is less than 1 per cent over the entire frequency range. Details of the panel assembly, control panel, and consolette were explained.

March 5, 1941, L. J. Black, chairman, presiding.

Seattle

Sidney Pickles presented his paper on "Antennas for C.A.A. Transoceanic Communication Systems" which is summarized in the report of the March 6 meeting of the Portland Section which appears in this issue.

February 28, 1941, K. H. Ellerbeck, chairman, presiding.

Toronto

A. B. Chamberlain, chief engineer of the Columbia Broadcasting System, presented a paper on the "CBS International Broadcast Facilities."

A review of the history of international broadcasting was first presented. This service commenced in 1924 and was known as experimental relay broadcasting. Coincident with this development in the United States, a parallel development occurred in Great Britain, Holland, Germany, France, and other European countries. Today there are almost 300 international broadcast stations in the world and approximately 200 of these are in South America. In the United States there are six licensees operating a total of twelve stations, all of which will be using 50 kilowatts by next year. The significance of this service under present world conditions was indicated.

The facilities of the Columbia Broadcasting System were then described. A pair of 50-kilowatt transmitters are being installed at Brentwood, Long Island, together with a frequency-modulation relay station to provide for program service from New York studios. These transmitters are located on the site of the Mackay Radio and Telegraph Company station.

There are thirteen directive antennas being constructed. The largest will be 220 feet high and 1100 feet long with 32 elements arranged in four sections. Any of the antennas may be coupled to either of the transmitters.

The largest antenna is directed toward Europe and will be used simultaneously by Columbia on 6120 kilocycles and by Mackay on 6935 kilocycles. Filters are used to prevent the output of one transmitter from being fed into the other transmitter. Experiments have shown that the system will operate satisfactorily with a frequency differential of only 5 per cent between the two transmissions.

March 24, 1941, G. J. Irwin, past chairman, presiding.

Washington

R. D. Wyckoff, staff geophysicist for the Gulf Research and Development Company, presented a paper on "Geophysical Exploration for Oil."

An outline was presented of the radio and geophysical principles and instruments used in the past and at the present time by the petroleum industries in their search for new oil reserves. The construction of various instruments used in determining various earth strata densities was described.

The construction and operation of the Gravinometer, by means of which formations thousands of feet below the surface of the earth and water may be determined, were described in detail. The methods of taking readings of gravitational effects accurate to within one part in ten million were shown together with the precautions necessary to avoid the effects of surrounding forces and to guard against false readings. A description was also given of methods used in determining strata formations by means of magnetic deflections and of the equipment and procedure used in making seismograph reverberation charts and contours.

The paper was concluded with a showing of several reels of motion pictures illustrating the difficulties encountered in the transportation of equipment to sites from which measurements must be made. They included operations in the United States, Norway, and Arabia.

March 10, 1941, M. H. Biser, chairman, presiding.

Errata

In the report on the January 13, 1941 meeting of the Pittsburgh Section which appears on page 38 of the January, 1941, PROCEEDINGS summarizing the paper "Dry Rectifiers" by L. O. Grondahl, of the Union Switch and Signal Company, the statement was made, "From 1920 through 1926, the copper-oxide rectifier came into common use and was replaced to a large extent starting in 1930 with the selenium rectifiers." The paragraph should read as follows:

"The history of dry rectifiers was traced from the introduction of the copper-sulphide type about 1906, through the period from 1920 to 1926 in which copper-oxide rectifiers came into common use, to the period between 1930 and the present in which selenium rectifiers have been introduced.

"The speaker also pointed out that the copper-sulphide rectifier and the selenium rectifier are of the nature of electrolytic rectifiers, since they require forming by an electric current to produce their rectifying characteristics. In this respect, the copper-oxide rectifier is in a distinct class since no forming is necessary and no change in the rectifier is produced by the passage of a current."

These errors occurred in the editing of the report submitted by the section secretary.

Membership

The following indicated admissions to membership have been approved by the Admissions Committee. Objections to any of these should reach the Institute office by not later than May 31, 1941.

Admission to Associate (A), Junior (J), and Student (S).

Anthony, A. R., (A) 834 S. Monroe St., Arlington, Va.
Antman, M. A., (A) c/o Ballantine Laboratories, Boonton, N. J.
Arndt, W. R., (S) 1527 Yale Station, New Haven, Conn.
Baker, J. H., Jr., (J) Crystal City, Texas
Bennett, R., (S) 155 Westminster Ave., Montreal West, Que., Canada
Blom, B. V., (A) 160 Fenimore St., Brooklyn, N. Y.
Brewer, A. H., Jr., (A) 314 S. Superior St., Angola, Ind.

Brown, C. W., (A) University of Idaho, Moscow, Idaho
Brown, E. W., (A) 2167 Broadway, Gary, Ind.
Brown, W. C., (S) 548 Prince Arthur St., W., Montreal, Que., Canada
Bruck, G. G., (A) 1624 Marlowe Ave., Cincinnati, Ohio
Carty, D. G., (S) 160 Waverley St., Ottawa, Ont., Canada
Castrignano, R. A., (J) 68 Thompson St., New York, N. Y.
Clewes, T. W., (A) 510 N. Wayne St., Angola, Ind.
Custer, H. M., (S) c/o P.O. Department, Johnstown, Pa.
Dale, D. L., (A) 386 Minnesota St., St. Paul, Minn.
Doll, H. G., (A) 2720 Leeland Ave., Houston, Texas
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Finney, R., (A) 42 Elm St., Elizabeth, N. J.
George, H. H., (S) 292 Lafayette Ave., Brooklyn, N. Y.
Grubb, G. R., (A) School of Technical R.A.F. Station, Ambala, Punjab, India
Guillemin, E. A., (A) Massachusetts Institute of Technology, Cambridge, Mass.
Hall, E. C., (A) 5327 Abbott Pl., Los Angeles, Calif.
Hobbs, C. F., (A) 300 N. 46th St., Belleville, Ill.
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Hudson, C. O., (A) Box 196, North Charleston, S. C.
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Kendall, P. R., (A) 1605 E. 73rd St., Cleveland, Ohio
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Lester, J. M., (A) c/o Sperry Gyroscope Co., Garden City, N. Y.
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Lloyd, P. A., (S) 1062 Adams St., Corvallis, Ore.
Mackenzie, D. G., (A) 2758 E. 16th Avenue, Vancouver, B. C., Canada
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Mellor, H. M., (A) 2 Princes Ave., Finchley, London, N. 3, England
Nason, D. B., (A) 1628 Marlowe Ave., Cincinnati, Ohio
Partington, G. E., (A) Marconi Research Labs., Great Baddow, Chelmsford, Essex, England
Pearson, R. V., (A) Radio Station WBML, Macon, Ga.
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Spiegel, S., (J) 3560 Rochambeau Ave., New York, N. Y.
Stacey, D. S., (S) 44 Follen St., Cambridge, Mass.
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Sulman, I., (A) Kibla St. 37/20, Basrah, Iraq
Swire, B. E., (S) 78 Arabella St., Longueville, N.S.W., Australia
Szetela, F. E., (A) 2211 Roslyn Ave., Baltimore, Md.
Thompson, D. E., (A) Howard Circle, Box 105, Emporium, Pa.
Van Baalen, J. M., (A) 1016 Amherst Ave., Buffalo, N. Y.
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Weed, A. C., Jr., (S) 119 S. Grant St., West Lafayette, Ind.
Whitney, S. D., (A) c/o Siebenthaler Div., Aircraft Accessories Corp., 410 W. 6th St., Kansas City, Mo.
Wirsu, O. L., (S) 31A Salisbury Rd., Kensington, N.S.W., Australia
Woodward, R. H., 19 Everett St., Cambridge, Mass.
Woodward, R. O., (A) 1333 E. Main St., Louisville, Ohio
Wooley, R. L., (S) M.I.T. Dormitories, Cambridge, Mass.
Yager, J. C., (A) Box 252, Chula Vista, Calif.
Zwaska, J. F., (S) 404 Touhy Ave., Park Ridge, Ill.

Books

The Radio Amateur's Handbook, Eighteenth (1941) Edition, by the Headquarters staff of the A.R.R.L.

Published by the American Radio Relay League, Inc., West Hartford, Conn. 552 pages, including 8-page topical index and 96-page catalog section of amateur radio equipment. Approximately 830 illustrations and 90 charts and tables. $6\frac{1}{2}\times 9\frac{1}{2}$ inches. Price, paper bound, $1.00 in continental U.S.A., $1.50 elsewhere; buckram bound, $2.50. Spanish edition, $1.50.

The Handbook deals particularly well with the three major constituents of a radio station, the receiver, the transmitter, and the antenna system. Considerable space is devoted to each, giving design factors and construction details for numerous practical examples. A tabulation of characteristics and miscellaneous data relative to over 600 types of vacuum tubes is included.

Other sections cover briefly such subjects as fundamental principles, regulations, station operations, workshop practices, League activities, etc.

The quality and format of publication is good and the Handbook should prove useful to anyone interested in amateur radio.

H. O. Peterson
R.C.A. Communications, Inc.
New York, N. Y.

Electromagnetic Devices, by Herbert C. Roters.

Published by John Wiley and Sons, Inc., 440 Fourth Ave., New York, N. Y. 561 pages, price $6.00.

This design manual, while not especially intended for radio engineers, is valuable to them because they use magnetic devices in the form of relays and other mechanisms for switching, keying, and remote control.

The treatment of this subject comprises a well-balanced combination of the fundamental principles and the practical problems of design. It is directed to graduates in electrical engineering. There is an excellent treatment of the forces involved in magnetic devices, followed by practical procedures for calculating the permeance of magnetic-flux paths. Special attention is given to the properties of available magnetic materials and coils. The problems treated in detail include tractive magnets, time-delayed magnets, high-speed magnets, alternating-current magnets, and relays. The generous number of specific examples and the completeness of their description makes this book especially useful as a reference.

Harold A. Wheeler
Hazeltine Service Corporation
Little Neck, L. I., N. Y.

Report of the Secretary—1940

This report on the activities of the Institute during 1940 is published for the information of the membership.

Membership

The paid membership increased 1.6 per cent during the year and totals 5705. The membership figures throughout the life of the Institute are plotted In Fig. 1.

The proportions of the membership in the United States and its possessions and in the rest of the world for the past five years appear in Table I.

In 1937 the highest percentage of foreign membership was recorded. Since then that proportion has decreased steadily and is now at its lowest value since 1931. Considered separately, the foreign membership has decreased during 1940 by 13.9 per cent while the domestic membership increased by 5.9 per cent.

The membership in the British Empire was 15 per cent smaller at the end of 1940 than at the end of 1939 and a reduction of 28 per cent was recorded for the European membership for the same period. A 15 per cent drop was also noted in the Japanese membership. Membership in the South American republics increased by 13 per cent.

New Members

During 1940, 855 individuals were elected to membership in the Institute, an increase of 8 per cent over the 790 who were admitted in 1939. About 6 per cent more applications for membership were received during 1940 than in 1939, the total being 913 as contrasted with 858. The number of Student applications received was about 14 per cent greater than for the previous year and for other grades the increase was approximately 2 per cent.

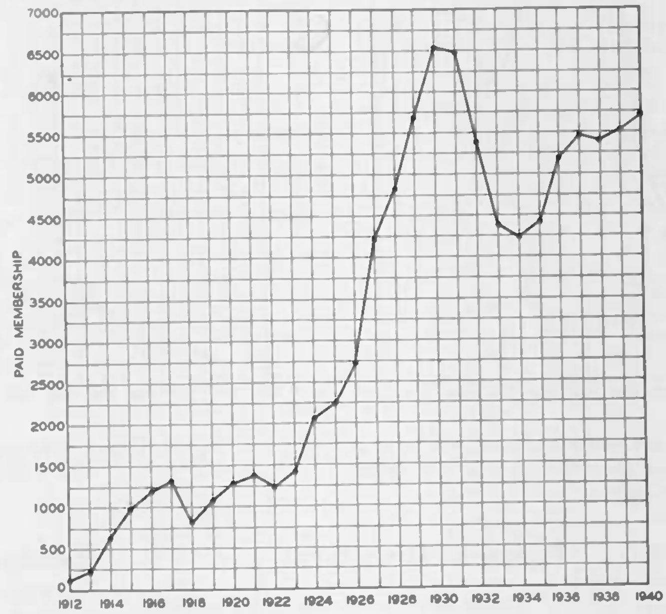

Fig. 1—The total number of paid members at the end of each year of the life of the Institute is plotted above.

Proceedings

Serious difficulties were encountered in publishing Volume 28 of the PROCEEDINGS. In the publication of the previous volume, delays of about a year between the receipt of a paper and its appearance in the PROCEEDINGS occurred. During the last half of the year an active program to reduce this delay was instituted and worked so effectively that when the December 1939 issue was distributed there were no additional manuscripts awaiting publication.

An insufficient number of manuscripts were approved for publication in time to permit the regular distribution of the February issue and this condition persisted throughout the year. As a result of the accumulated delay, the December issue was not published until mid-February, 1941.

Volume 28, which contained 586 pages of technical and Institute material (numbered in arabic), was substantially smaller than any volume issued during the last dozen years. This reduction was caused entirely by a lack of available acceptable material; a substantial sum of money provided in the budget for printing was not expended.

Distribution by Grades

A breakdown of the membership by grades is given in Table II. The Fellow, Member, and Student grades show increases both numerically and as proportions of the total membership. The Associate grade shows a reduction of about 1 per cent over last year and the Junior grade accounts for a smaller fraction of 1 per cent than was previously the case.

TABLE I

DOMESTIC AND FOREIGN MEMBERSHIP

	1936	1937	1938	1939	1940
Total	5196	5459	5403	5612	5705
United States and Possessions	3988	4145	4126	4408	4668
Foreign	1208	1314	1277	1204	1037
Per Cent Foreign	23.2	24.1	23.6	21.5	18.2

TABLE II

MEMBERSHIP DISTRIBUTION BY GRADES

	1936	1937	1938	1939	1940	Per Cent 1940
Fellow	133	136	156	159	175	3.1
Member	637	624	645	660	700	12.3
Associate	4093	4291	4250	4362	4314	75.5
Junior	34	48	38	32	20	0.4
Student	299	360	314	399	496	8.7
	5196	5459	5403	5612	5705	100.00

Sections

Some idea of the meetings activities and membership of our 23 sections will be obtained from Table III. The 200 meetings

TABLE III

SECTION MEMBERSHIP AND MEETINGS

	Membership Dec. 31, 1940	Meetings Held			1940 Average Attendance[1]	1940 Per Cent Attendance
		1938	1939	1940		
Atlanta	35	11	10	10	23	66
Baltimore[2]	82	—	2	8	90	110
Boston	231	5	9	3	134	58
Buenos Aires[2]	54	—	2	9	43	80
Buffalo-Niagara	49	10	3	10	63	129
Chicago	296	14	9	9	187	63
Cincinnati	89	10	10	10	49	55
Cleveland	78	6	11	4	35	45
Connecticut Valley	84	6	4	10	42	50
Detroit	104	10	10	10	72	69
Emporium	90	12	12	13	60	67
Indianapolis	62	8	3	1	32	52
Los Angeles[3]	224	11	10	15	119	53
Montreal	71	10	7	10	50	70
New Orleans	11	3	2	—	—	—
Philadelphia	317	9	8	8	201	63
Pittsburgh	58	11	12	10	30	52
Portland	56	1	11	9	43	77
Rochester[4]	36	10	11	12	—	—
San Francisco	214	17	12	14	62	29
Seattle	63	9	8	10	69	109
Toronto	79	11	9	5	87	110
Washington	256	10	10	10	115	45
	2639	194	185	200		

[1] Does not include joint meetings with other societies.
[2] Established October, 1939.
[3] Seven meetings credited for 1940 Pacific Coast Convention.
[4] Six meetings credited for 1940 Rochester Fall Meeting. All meetings held jointly with other societies.

is a new high value for the life of the Institute. The number of members residing in section territory was exceeded only in 1930 when our peak total membership occurred. The 1930 figures gave 42 per cent of the membership in sections against 46 per cent for 1940. In neither case have those residing in and near New York City been included.

was devoted to informal discussions of advanced work in the electronics field. We are indebted to Stevens Institute of Technology for the use of their facilities for the meeting. The registration totaled 252.

On November 11, 12, and 13, the twelfth Rochester Fall Meeting was held. There were 20 technical presentations. The attendance was 517.

Finances

At the end of this report will be found a balance sheet and a statement of income and expenses abstracted from our annual audit prepared by Patterson & Ridgway, Certified Public Accountants. In addition, the income and expenses over the life of the Institute are plotted in Fig. 3.

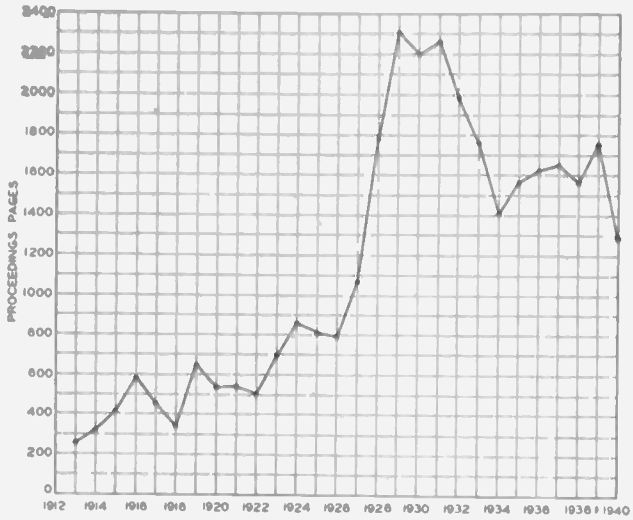

Fig. 2—Pages of technical and Institute material (numbered in arabic) published each year are shown graphically. The new larger page is equivalent to 2.2 smaller pages.

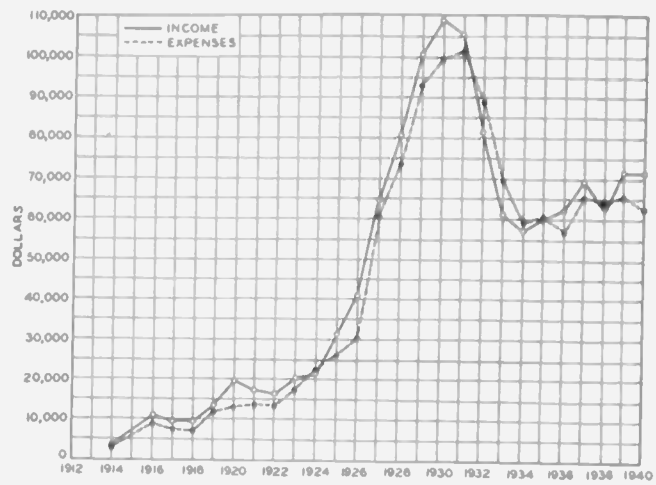

Fig. 3—Income and expenses are plotted for the life of the Institute.

Meetings

In addition to the 200 meetings held by our sections, there were 8 meetings in New York City and 5 of the convention type which are noted below.

The annual meeting of the American Section of the International Scientific Radio Union (U.R.S.I.) and the Institute occurred on April 26 in Washington, D. C. There were 17 papers presented. The attendance was approximately 125.

The fifteenth Annual Convention was held on June 27, 28, and 29 in Boston, Massachusetts. Twenty-four technical papers were presented. There were 9 trips scheduled, one of which was canceled because of weather conditions. Three of the trips were primarily for women. The attendance totaled 955 men and 116 women.

The fourth Pacific Coast Convention was held in Los Angeles, California, on August 28, 29, and 30. In addition to the presentation of 23 papers, there were two informal seminar discussions. Four trips were included in the program. Three hundred and fifty-two men and 24 women were in attendance.

On October 11 and 12, the third Electronics Conference was held. The meeting

Board of Directors

Eleven meetings of the Board of Directors were held during 1940. The 21 members on the Board are responsible for the management of the Institute.

Committees

Many of the Institute's activities are directed or carried out through committees of which 20 are of the administrative type and 33 are technical. These committees held 128 meetings during 1940.

Awards

At the Annual Banquet which was held in Boston on June 28, the Medal of Honor was presented to Lloyd Espenschief for his accomplishments as an engineer, an inventor, a pioneer in the development of radiotelephony, and for his effective contribution to the progress of international radio coordination.

At the same function, the Morris Liebmann Memorial Prize was awarded to Harold Alden Wheeler for his contribution to the analysis of wide-band high-frequency circuits particularly suitable for television.

Headquarters Staff

Of the eleven members of the headquarters staff, six have served the Institute for ten or more years.

Deaths

The deaths of one Fellow, one Member, and six Associates were reported during 1940. Their names are given below:

Bradley, R. A.	Long, J. J.
Carson, J. R.	Murphy, F. M. G.
Heyes, Oswald	Rangaswamiengar, K. S.
Kellogg, L. A.	Sandy, B. G., Jr.

Acknowledgment

To those members who give of their time and energies to participate in the management and operation of the Institute and its sections, a sincere thanks is due. Without their efforts the Institute would be a much less effective organization.

Respectfully submitted,

Harold P. Westman

HAROLD P. WESTMAN
Secretary

February 27, 1941

Comparative Statement of Income and Expenses for the Years Ending December 31, 1940 and 1939

			1940	1939
INCOME				
Dues, Current and in Arrears			$40,487.50	$39,538.75
Entrance and Transfer Fees			1,774.00	2,023.00
Subscriptions			11,399.95	12,816.70
Advertising			13,138.66	9,880.24
Binders, Bound Volumes, Emblems, and Reprints			1,795.55	2,857.64
Interest from Investments; Including Morris Liebmann Memorial Fund			916.00	774.90
Conventions			1,770.00	3,470.00
Miscellaneous			502.87	377.93
Total Income			$71,784.53	$71,739.16
EXPENSES				
Advertising			$ 66.73	$ 325.56
Awards			341.00	341.00
Bad Debts Written Off, Less Recoveries*			4,874.42	3,342.19
Binders, Bound Volumes, Emblems, Reprints			1,380.64	2,440.16
Conventions			3,023.12	2,835.46
New York Meetings			796.22	985.28
Office			4,128.68	3,725.12
	1940	1939		
Depreciation of Furniture and Fixtures	$ 449.94	$ 447.72		
Insurance	180.45	143.52		
Postage	1,543.92	1,530.71		
Stationery and Supplies	1,240.15	931.76		
Telegraph and Telephone	714.22	671.41		
Printing			17,255.23	20,996.78
Proceedings	16,829.27	20,779.35		
Standards	88.30	16.61		
Yearbook	—	—		
Miscellaneous	337.66	200.82		
Rent and Electricity			3,228.00	3,225.15
Salaries			22,560.20	22,141.04
Advertising	1,113.50	3,465.00		
General	13,832.94	9,657.22		
Proceedings	6,041.06	7,137.81		
Standards	1,572.70	1,881.01		
Sections			2,987.69	3,078.38
Miscellaneous			2,069.02	1,984.16
Total Expenses			$62,710.95	$65,420.28
Addition to Reserves			$ 9,073.58	$ 6,318.88
			$71,784.53	$71,739.16

* These figures cover chiefly nonpayment of dues.

The Institute of Radio Engineers, Inc.
Comparative Balance Sheet

December 31, 1940 and 1939

Assets	December 31, 1940	December 31, 1939	Increase **Decrease**
Current Assets			
Cash	$33,131.01	$21,230.43	$11,900.58
Accounts Receivable—Current			
Dues	450.72	489.52	**38.80**
Advertising	3,153.00	702.66	2,450.34
Reprints	51.17	146.71	95.54
Inventory (As submitted by the management)			
Proceedings	6,901.84	6,912.56	**10.72**
Bound Volumes	194.00	202.00	**8.00**
Binders	278.45	152.15	126.30
Emblems	360.34	275.58	84.76
Accrued Interest on Investments	167.50	185.83	**18.33**
Total Current Assets	$44,688.03	$30,297.44	$14,390.59
Investments—At Cost			
Securities Owned by the Institute	36,947.87	49,922.12	**12,974.25**
Morris Liebmann Memorial Fund	10,012.45		10,012.45
Total Investments—at Cost (Market Value—12/31/40—$27,773.82)	$46,960.32	$49,922.12	**$ 2,961.80**
Furniture and Fixtures after Reserve for Depreciation	3,289.30	2,245.00	1,044.30
Prepaid Expenses			
Unexpired Insurance	85.12	87.64	**2.52**
Stationery Inventory—Estimated	200.00	200.00	
Convention Expense	475.95	65.86	410.09
Total Assets	$95,698.72	$82,818.06	$12,880.66

Liabilities and Surplus	December 31, 1940	December 31, 1939	Increase **Decrease**
Accrued Wages	$ 345.88		$ 345.88
Accounts Payable	4,220.46	$ 252.61	3,967.85
Section Rebate—Buenos Aires	44.50		44.50
Suspense	15.07	24.07	**9.00**
Advance Payments			
Dues	2,119.16	2,966.06	**846.90**
Subscriptions	2,735.37	2,571.38	163.99
New York State Income Tax Withheld from Employees	50.56	38.00	12.56
Total Liabilities	$ 9,531.00	$ 5,852.12	$ 3,678.88
Funds			
Morris Liebmann Memorial Fund	$10,012.45	$10,000.00	$ 12.45
Associated Radio Manufacturers Fund		1,997.80	**1,997.80**
Total Funds	$10,012.45	$11,997.80	**1,985.35**
Unexpended Income			
Morris Liebmann Memorial Fund	$ 77.87	$ 77.87	
Deferred Income—Convention 1941	615.00		$ 615.00
Surplus—Earned	73,464.60	64,890.27	8,574.33
Surplus—Donated			
Transfer of Associated Radio Manufacturers Fund—Authorized by Board of Directors—March 6, 1940	1,997.80		1,997.80
Total Liabilities and Surplus	$95,698.72	$82,818.06	$12,880.66

Contributors

Robert R. Buss (S'37) was born on March 14, 1913, at Provo, Utah. He received the A.B. degree in mathematics from San Jose College in 1935. From 1935 to 1939 he was a Newell Scholar at Stanford University, receiving the E.E. degree

Robert R. Buss

in 1938 and the Ph.D. degree in 1940. Dr. Buss was a laboratory assistant in electrical engineering at Stanford University during 1936 and 1937 and an engineer at Heintz and Kaufman, Ltd., during 1939 and 1940. Since 1940 he has been an engineer at the Litton Engineering Laboratories. He is a member of Tau Beta Pi and the Society of Sigma Xi.

❖

Bernard D. Loughlin (A'40) was born in New York City on May 19, 1917. He received the B.E.E. degree in electrical engineering from Cooper Union Institute of

Bernard D. Loughlin

Technology in 1939. While an undergraduate, he developed a phase curve indicator and presented a paper on the device before the Student Convention of the American Institute of Electrical Engineers, winning a prize. Since graduation, he has been employed by the Hazeltine Service Corporation doing television research work, including the development of a phase curve tracer.

❖

Chao-Ying Meng* was born in the Lao-Ting district, Hopei Province, China, on November 9, 1906. He received the B.S. degree in physics in 1928 and the M.S. de-

Chao-Ying Meng

gree in 1931 from the Yenching University, Peiping, China, and the Ph.D. degree in physics from the California Institute of Technology in 1936. From 1936 to 1937 he was a lecturer in the physics department of Yenching University. Dr. Meng joined the Tsinghua University in the summer of 1937 and is now a professor of the Radio Research Institute there.

* Paper appeared in the December, 1940, issue of the Proceedings.

❖

Hans Salinger

Hans Salinger (A'37) was born in Berlin, Germany, on April 1, 1891. He received the Ph.D. degree from the University of Berlin in 1915. From 1919 to 1929 he was a research associate at the Reichpostzentralamt in Berlin and from

Frederick E. Terman

1929 to 1935, professor at the Polytechnical Institute and the Heinrich Hertz Institut für Schwingungsforschung in Berlin. Since 1936 Dr. Salinger has been with the Farnsworth Television and Radio Corporation. He is an Alumni Member, University of Pennsylvania Chapter, of Sigma Xi.

❖

Frederick Emmons Terman (A'25–F'37) was born on June 7, 1900, at English, Indiana. He received the A.B. degree in 1920 and the degree of Engineer in 1922 from Stanford University, and the Sc.D.

Alexander H. Wing

degree from Massachusetts Institute of Technology in 1924. From 1925 to 1937 Dr. Terman was an instructor, assistant professor, and associate professor of elec-

trical engineering at Stanford University. Since 1937 he has been professor and head of the electrical engineering department at Stanford. Dr. Terman was Vice President of the Institute of Radio Engineers in 1940 and President in 1941.

❖

Alexander H. Wing (A'39) was born in Yonkers, N. Y., on December 27, 1905. From 1923 to 1929 he was a student in Columbia College and in the School of Engineering (then the Schools of Mines, Engineering, and Chemistry) of Columbia University, receiving the degrees A.B. in 1927, B.S. in 1928, and E.E. in 1929.

From 1929 to 1932 he was in the employ of the General Electric Company at Schenectady, New York, and at West Lynn, Massachusetts. Since 1932 he has been an instructor in electrical engineering at the School of Technology of the College of the City of New York. In 1936 he was licensed as a professional engineer by the University of the State of New York.

Mr. Wing was elected to membership in the national honorary societies of Phi Beta Kappa, Tau Beta Pi, and Sigma Xi. He is an Associate member of the American Institute of Electrical Engineers, and a Member of the Society for the Promotion of Engineering Education.

Each Can Say

"I WAS A CLERK"

"I WAS A LINEMAN"

"I WAS A DRAFTSMAN"

Thirty-seven years ago, in 1904, the president of the American Telephone and Telegraph Company went to work as a clerk in one of the Bell System companies.

About that time, the 18 men who are now the presidents of the Bell telephone companies were starting their careers. For, like the head of the System, they have worked many years in the business — an average of 38 years each. Each of them can say: "I was a clerk," "I was a lineman," "I was a draftsman"— and so on.

The "know how" is here — for the every-day job of running the telephone business or to serve you in emergency. Up-from-the-ranks management is doubly important these days.

THE BELL SYSTEM IS DOING ITS PART IN THE COUNTRY'S PROGRAM OF NATIONAL DEFENSE.

BELL TELEPHONE SYSTEM

Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.
A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

* DIE AUSBREITUNG DER ELEKTROMAGNETISCHEN WELLEN (The Propagation of Electromagnetic Waves). BY BRUNO BECKMANN. Leipzig: Akademische Verlagsgesellschaft M.B.H., 1940. x+271+11 index pages, illustrated, 6½×9 inches, cloth, 25.60 rm.; paper 24 rm.

* ELECTROMAGNETIC DEVICES. BY HERBERT C. ROTERS, Director of Research, Fairchild Aviation Corporation. New York: John Wiley & Sons, Inc., February, 1941. 561 pages, illustrated, 6×9 inches, cloth. $6.00.

* ELECTRON-INERTIA EFFECTS. BY F. B. LLEWELLYN, Bell Telephone Laboratories, Inc. New York: The Macmillan Company, April, 1941. x+102+2 index pages, illustrated, 5½×8½ inches, heavy paper. $1.75.

* ELEKTRONENROHREN ALS END UND SENDERVERSTARKER (Electron Tubes as Power and Transmitter Amplifiers). BY HORST ROTHE and WERNER KLEEN. Leipzig: Akademische Verlagsgesellschaft M.B.H., 1940. x+137+4 index pages, illustrated, 6½×9 inches, cloth, 11 rm.; paper 9.40 rm.

POSITIONS OPEN

•

The following positions of interest to I.R.E. members have been reported as open on April 18. Make your application in writing and address to the company mentioned or to

Box No. ......
PROCEEDINGS of the I.R.E.
330 West 42nd Street, New York, N.Y.

Please be sure that the envelope carries your name and address

JUNIOR ENGINEER

A firm of consulting engineers engaged in making surveys for broadcasting stations has an opening for a young engineer. He will be expected to make field-intensity surveys and do other field work. A recent college graduate with the necessary mathematical and engineering background for this type of work is preferred. Box 240.

SALES ENGINEER

A large manufacturer wants a young engineer with the necessary technical and personal qualifications to engage in contact work with radio manufacturers. He should have an electrical engineering education, preferably with specialization in radio, and be interested in sales engineering. Box 241.

(Continued on page iv)

DAVEN
Precision LEVEL INDICATOR

The Type 910 Volume Level Indicator is an audio level indicator designed for service in broadcasting, sound recording, and allied fields. The meter is sensitive, rugged and correctly damped for program monitoring. The meter multiplier is a heavy duty, step type "T" attenuator designed to offer a constant impedance both to the line, and to the meter at all steps of control. The zero adjustment provides corrections of ± 0.5 Db. in 0.1 Db. steps. The input impedance of the 910 is 7500 ohms. The reference level is 1 mw. into 600 ohms.

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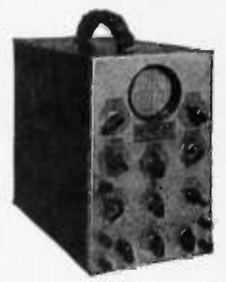
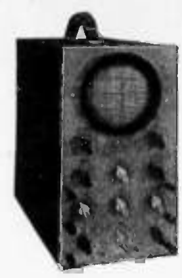
Select the
OSCILLOGRAPH
that best fits
YOUR JOB

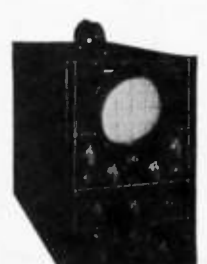
★ Here are three of the several types of DuMont cathode-ray oscillographs that give you a wide choice in selecting that instrument best fitted to your needs:



Type 164: An exceptionally compact, portable and inexpensive 3-inch instrument, ideally suited to production testing. Incorporates a vertical amplifier with voltage gain of 70 times, and a horizontal amplifier with voltage gain of 40 times, over the frequency range from 18 to 30,000 sinusoidal cycles per second. Horizontal amplifier amplifies either the sweep circuit or any external signal. Deflection plate terminals externally available.



Type 168: Moderately-priced 5-inch oscillograph. Larger screen size makes it ideal for studies requiring greater definition, particularly for lecture demonstration work. Vertical amplifier has voltage gain of 450 times. Input signal may be connected to either the amplifier input or to deflection plates, at front panel. Sweep circuit has a repetition rate of 15-30,000 times per second.



Type 208: Incorporates many desirable improvements and refinements, making it ideal laboratory oscillograph. Utilizes the DuMont intensifier-type cathode-ray tube with four deflection plates for use with balanced signal circuits. Vertical amplifier has voltage gain of 2000 times over a frequency range of 2 to 100,000 sinusoidal cycles per second. Linear time base has a repetition rate variable from 2 to 50,000 times per second. Instantaneous position control. Regulated power supply.

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★ Information on these and other DuMont Cathode-Ray Oscillographs, Cathode-Ray Tubes, Electronic Switch and associated electronic equipment, contained in Catalog B. Copy sent on request written on your business letterhead.

DUMONT
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Aerovox Type AF Prong-Base Electrolytic, alongside match box for size comparison. Also metal and bakelite mounting washers. In 1" and 1⅜" dia. cans, 25 to 450 v. D.C.W., and in capacities from 10 to 80 mfd., and higher to order. Various combinations to meet any requirements.

The PRONG-BASE ELECTROLYTIC *that is different...*

● Similar in appearance and purpose to the conventional prong-base electrolytics in general use today, the recently-introduced Aerovox Type AF incorporates several vital refinements in making this type still more popular with designers, manufacturers, servicemen and equipment owners.

The heretofore decidedly wobbly terminals are now rigid. The danger of shearing cathode tabs passing between can shoulder and cap or plug, is entirely eliminated. Electrolyte leakage and attendant corrosion troubles are done away with. Proper venting relieves any excess gas pressures. All in all, Aerovox has made the prong-base electrolytic a practical, safe, still more popular condenser for today's assemblies and servicing.

Submit Your Problem . . .

● Whatever your capacitance problem may be, send it along for our engineering collaboration. And if you can use the prong-base type, write us for engineering data on our AF improved construction. Samples, specifications, quotations, cheerfully submitted to responsible parties.

AF CONSTRUCTION

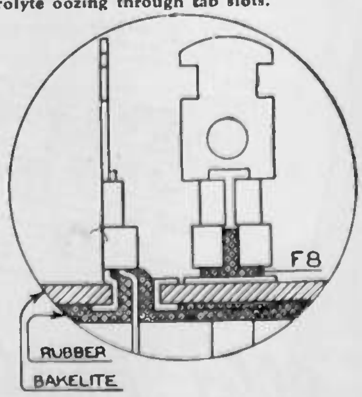
● First point of departure from conventional design is the square can shoulder in place of usual 30° slope. This permits cap or plug to seat solidly in place, making the seal more positive, and eliminating danger of shearing cathode-tab.

In place of usual two bakelite discs separated by sheet of flat rubber, AF construction employs a cup-shaped molded soft-rubber disc with side walls. The single bakelite disc fits within the walls of the cup-shaped disc which in turn rests squarely on flat shoulder of can.

Cup-shaped rubber disc has several slotted protrusions or sleeves molded in same. Through said sleeves pass the anode or positive tabs which, beyond the bend inside of sleeve, join with soldering lugs (see diagram below). This construction assures a rubber-sealed terminal. Electrolyte cannot reach the junction of tab and lug.

The lugs are actually eyeletted to bakelite disc —not just held between discs. All strain is removed from anode tabs. Impossible to loosen connections. Cathode tab is spot-welded to mounting ring.

Positive pin-hole vent instantly responsive to excess gas pressures yet normally self sealing. This in contrast to usual construction with gases and electrolyte oozing through tab slots.



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WORLD'S GREATEST all around ELECTRIC TOOL

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The new WHIZ ELECTRIC TOOL is the handiest power tool ever made. A rugged tool for power and precision work. Drills through ¼ inch iron plate in 42 seconds or engraves intricate designs. Handles any material: Metals—Woods—Alloys—Plastics—Glass—Steel—etc. Saves time. Eliminates labor. Plug into any socket AC or DC, 110 volts. Chuck ¼ inch capacity. Ball bearing thrust. Powerful, triple-geared motor STANDARD MODEL, with Normal Speed (uses 200 different accessories, instantly interchangeable). Price only $7.95.

The only DRILL-TOOL with a full year's guarantee.

FREE Accessory outfit (Value $2) includes set of drills, mounted 1½ inch grinder, sanding discs, cutting wheels, mounted brush, polishing wheel, carving burr, etc. FREE with each tool ordered NOW. We pay postage.

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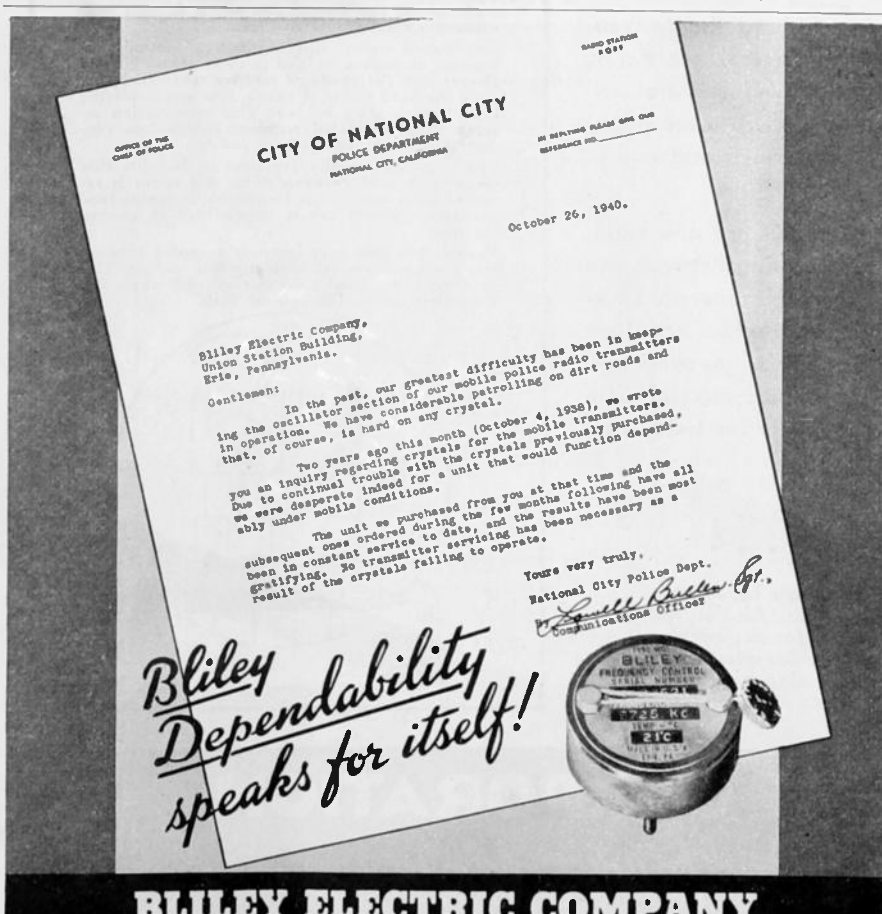


BLILEY ELECTRIC COMPANY
UNION STATION BUILDING ERIE, PA.

POSITIONS OPEN

(Continued from page ii)

RADIO ENGINEERS

A new drive has been begun by the Civil Service Commission with the recent announcing of an Engineer examination which includes all branches of engineering except chemical, metallurgical, marine, and naval architecture for which examinations have been announced previously.

Applicants must have completed a 4-year recognized college course, except that provision is made for the complete substitution of qualifying professional engineering experience of the proper type, quantity, and quality for the education lacking. Additional professional engineering experience, differing in kind, length, degree of progression and responsibility, according to the grade and branch of the position is applied for is also required except that graduate study in engineering may be substituted for part of the experience. The maximum age limit is 60 years.

Engineers qualified in certain specialized fields, including radio, are particularly needed for the National Defense program and are urged to file their application at once.

The duties of these positions will include design, construction, and research. The positions pay from $2,600 to $5,600 a year. Further information and application forms may be obtained at any first- or second-class post office or from the Civil Service Commission, Washington, D.C.

GEOPHYSICAL DEVELOPMENT

The geophysical laboratory of a well-known company has openings on its research staff for two or three men with training in electrical engineering and physics. Applicants should have had at least one year of graduate work, with emphasis on the non-communication applications of vacuum tubes, such as very low frequency amplifiers and non-linear control circuits. Box 242

SALES MANAGER

Opening for an aggressive sales manager with a progressive and expanding manufacturer of industrial and electronic equipment. Salary basis. Box 239.

ENGINEERING ASSISTANT

There is an opening in an eastern manufacturer's laboratory for an engineering assistant with some college-engineering education and with some familiarity with high-power test-instrument construction for radio transmitters up to 1 kilowatt. Ability to pursue development of electronic control circuits desirable. Present staff knows of this opening. Box 238.

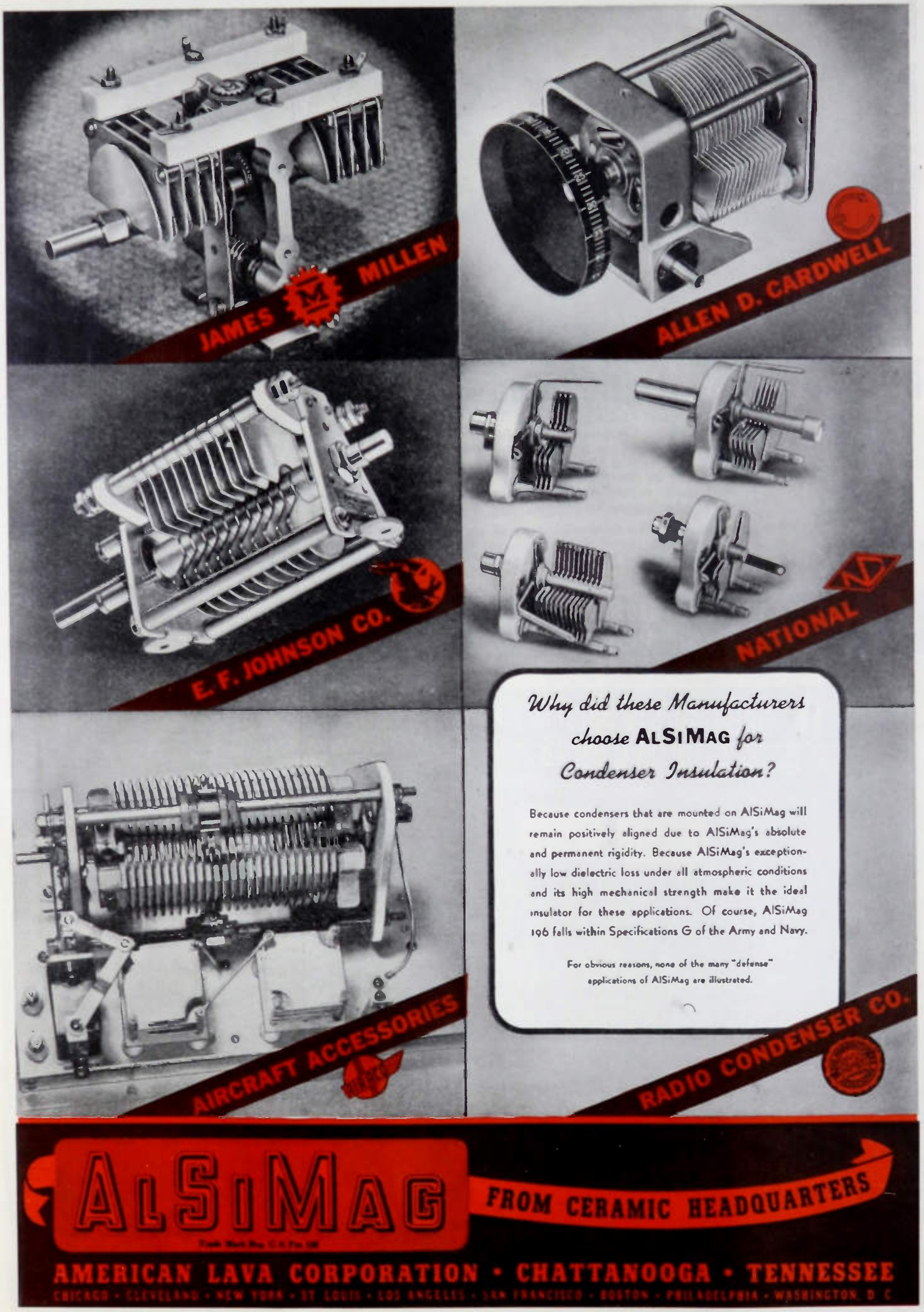
RADIO ENGINEERS

An expanding manufacturer needs engineers for development and design work. Should be engineering graduates and have had television and goniometry experience. Liberal salaries for men who can produce. Openings for a junior a senior and an executive. Present staff knows of these openings. Box 243

Attention Employers . . .

Announcements for "Positions Open" are accepted without charge from employers offering salaried employment of engineering grade to I.R.E. members. Please supply complete information and indicate which details should be treated as confidential. Address: "POSITIONS OPEN," Institute of Radio Engineers, 330 West 42nd Street, New York, N.Y.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal



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NATIONAL
AIRCRAFT ACCESSORIES
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Measurements with the 301-A instrument have been simplified—it's nearly as easy to use as a standard broadcast field meter, and arranged for recording without additional amplifiers. With the 302-A noise meter attachment, surveys of signal to noise ratio may also be made. Leaders in UHF development and prominent consultants employ the RCA 301-A.

The 301-A operates on the same principle as broadcast instruments. Arranged primarily for amplitude modulation stations, it may be modified for measuring FM stations during program transmission simply by changing a resistor and condenser...or used without change to check unmodulated carrier. Meter measures only 9¼" x 13" x 20⅜", weighs 38 lbs. Accessory case contains doublet antenna and supporting tripod.

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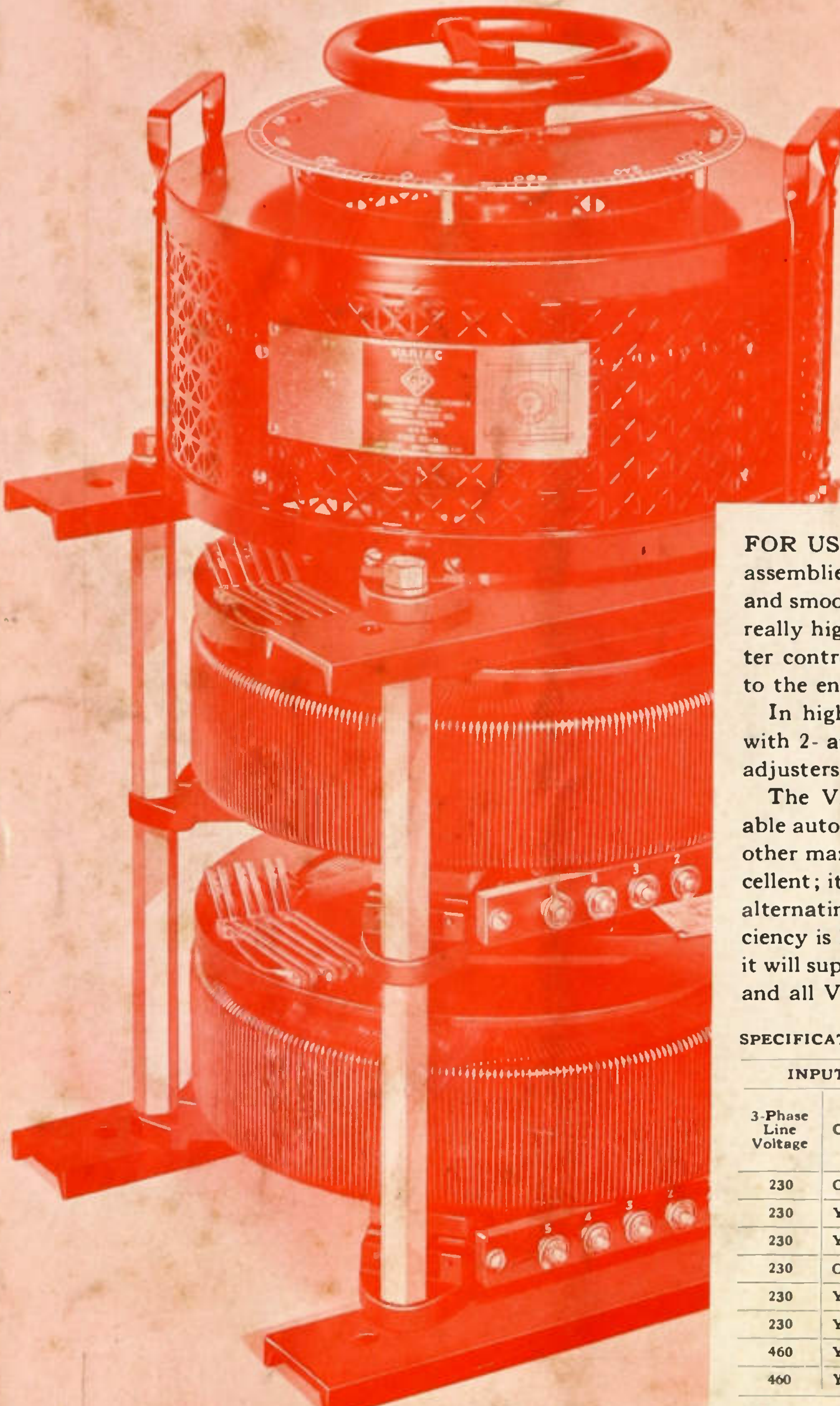
VARIACS



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SPECIFICATIONS FOR 2- AND 3-PHASE VARIAC COMBINATIONS

INPUT		OUTPUT			Type of Assembly	Price
		KVA				
3-Phase Line Voltage	Circuit	At Input Voltage	At Max. Voltage	3-Phase Line Voltage		
230	Open Δ	3.6	4.2	0–270	100-RG2	$ 85.00
230	Y	3.6	3.6	0–460*	100-RG3	130.00
230	Y	7.2	7.0	0–270	100-QG3	130.00
230	Open Δ	12.5	9.3	0–270	50-BG2	225.00
230	Y	12.5	8.0	0–460*	50-BG3	335.00
230	Y	18.	17.5	0–270	50-AG3	335.00
460	Y	7.2	7.2	0–460	100-RG3	130.00
460	Y	25.0	25.0	0–460	50-BG3	335.00

* Open-circuit voltage—regulation is poor for this connection

Write for the NEW Variac Bulletin 688

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